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Soldering and Solderability of Electronic Components

R. Barrie
Mullard Central Technical Services

This article is intended to familiarise the reader with the problems that may arise when mechanised soldering methods are used. It also outlines procedures by which a satisfactory soldering standard for terminal connections may be ensured.

INTRODUCTION

The art of joining two metallic components by soldering is widely practised and has been in existence for a considerable number of years. It requires no special skills, and operators for hand soldering can be trained to solder efficiently in a relatively short time.

The same soldering methods which have been in use for many years are still employed today, with minor improvements, for both production and repair work. However, in recent years, mechanical methods of soldering have been devised and these are used by some manufacturers where they are particularly suitable to the assembly under construction. For example, where a printed panel is being used, the manufacturer may decide that a fully automated process is the quickest and the most economical way of soldering the many component parts to the printed circuit. Another manufacturer may still employ hand soldering for a similar piece of work, so that both hand and mechanical soldering methods may be seen in use in most assembly areas.

The end product of soldering, irrespective of the method, is the finished joint and therefore the only criterion used in soldering is whether the joint is satisfactory. Both hand and mechanical methods result in adequate joints being formed, provided that suitable controls are maintained. To ensure that a given standard of solderability is reached, there are three types of test currently in use for solderability in the United Kingdom; one for hand, two for mechanical methods, all of which are outlined later in this article.

The use of mechanical soldering necessitates careful consideration of solderability. The choice of a particular terminal wire and its finish is more critical for a mechanical process than for hand soldering and selection is determined by its solderability in a given time. The time factor is extremely important, since the period a component can safely be subjected to the thermal stress resulting from a mass of molten solder applied to its terminal wires is limited. Published data for devices indicate maximum permissible temperatures, and the duration of the maxima, at which they may be soldered.

With hand soldering processes, heatsinks are generally recommended for use on component terminal wires between the junction and the body of the device, and the hot solder bit is applied to the joints only. With the use of a mechanical soldering process, heatsinks are not used and an assembly panel may be subjected to several separate heat cycles which tend to be sustained by the thermal inertia of the panel itself. Thus components mounted on a panel reach higher temperatures with a mechanical process than components which are hand soldered. To overcome the risk of thermal damage to components during mechanical soldering, the time spent in each successive area of high temperature must be kept to a minimum. For this reason, the solderability standard of component terminal wires intended for flow or dip soldering must be higher than for hand soldering.

The two solderability test methods which have been used in the electronics industry for some years are:

(i) solder bit test,
(ii) solder bath test.

Recently, a further test method – originated by J. A. ten Duis (Ref. 1) – has been investigated by the Electronic Engineering Association, modified, and adopted as a method of testing round terminal wires without severance from the body of the component (Ref. 2). The International Electrotechnical Commission (Technical Committee 50) had been seeking a new, precise method of solderability assessment for some time, and a working group, which included J. A. ten Duis, carried out experiments on the modified equipment. These experiments resulted in further detailed modifications to the test equipment, and the I.E.C. has put forward proposals to the effect that the equipment is accepted as a method of determining the relative solderability of round terminal wires. Because the main interest of the electronics industry is in terminal wires, tags, spade terminals, etc., this article does not discuss the soldering of other, larger metal components.

SOLDERING

Soldering is a method whereby two or more wires may be permanently connected together with the use of heat and a solder alloy so that metallic – and therefore electrical – continuity is established.

The solder alloy becomes bonded to the surface of each wire, fills the joint area between wires, and acts as a link between them. Thus there are two solder bonds in each soldered joint as shown in Fig. 1. When an intermetallic compound occurs, a permanent unbroken solder film is formed and the solder alloy is said to have wetted the joint.

The solder alloy which is used for a joining process always has a melting point which is lower than those of the wires to be joined. Therefore, when heat is applied, the temperature of the joint area rises until it is high enough to melt the applied solder, the wires remaining solid while the solder wets the joint.

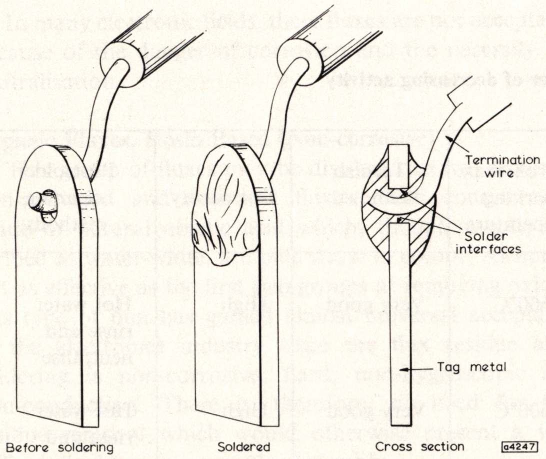

Fig. 1—Diagram of soldered tag to show solder-to-metal interfaces.

It is essential that the two wires to be joined are clean and free from contamination since any surface deposit or oxide will prevent the solder from wetting a joint. Thorough cleaning of a piece of wire intended for soldering is required, but since it is generally exposed in air during soldering, an oxide can form in this time. This problem is overcome by the use of a flux, as discussed in the next section.

FLUXES

In the making of a soft-soldered joint, fluxes have three important functions to perform; these are, heat transfer, tarnish removal and prevention of re-tarnishing.

To make a joint, both the joint members and the solder must be brought quickly to the temperature at which they will bond. Without the presence of a flux to ensure good thermal contact between the heat source (such as a soldering iron or solder bath) and the joint members, too much time may be spent waiting for the parts to heat, and neighbouring materials or components may become damaged.

Both the joint members and the solder have on their surfaces a film of tarnish, usually composed of metal oxides, sulphides or, more rarely, chlorides. These contaminations prevent the flow of molten solder over the joint member surfaces and must therefore be removed before bonding can occur. The flux dissolves these substances, leaving a clean metal surface to which the solder can adhere.

At normal soldering temperatures, metal surfaces oxidise very rapidly. After dissolving surface tarnishes, therefore, the flux must stop atmospheric oxygen from reaching the newly-cleaned metal surfaces.

The chemical reaction between a flux and tarnish can take several forms, including

(i) dissolving the oxide, with the formation of a salt and water
(ii) reducing the oxide to the metal
(iii) electroless plating.

Fluxes which employ the first reaction are the most common and are widely used in the electronics industry. Since they convert an oxide to salt and water, they must themselves be acid; in fact, they either consist of a free weak acid, or they contain a material which liberates a small quantity of a strong acid.

Reducing fluxes are generally gaseous; they remove the oxygen from metal oxides, leaving a film of finely divided metal powder. Electroless-plating fluxes (commonly called reaction fluxes) leave a thin film of metal on the joint member, much in the same way as copper sulphate solution will leave a film of copper on steel. This film is readily wetted by the solder.

Fluxes generally decompose and lose their fluxing ability when heated. Provided they retain their activity long enough to make the joint, this defect is unimportant and may even be an advantage in that the flux residue is thereby rendered less corrosive.

Fluxes are often mixed with a material known as a vehicle, the function of which is either to retain a volatile flux in a liquid state long enough for it to do its work of cleaning; or to reduce the temperature at which it becomes liquid and so improve its heat transfer characteristics. When used for the latter purpose the vehicle often takes the form of a volatile solvent such as alcohol, when additional heat will then have to be supplied to evaporate the solvent before the flux can commence to work.

Different types of fluxes are used for various soldering processes and they can be classified into three groups: highly corrosive, corrosive and non-corrosive. Details of each group are given separately below and a summary of each group is given in Table 1.

Inorganic Fluxes (Highly Corrosive)

While inorganic corrosive fluxes are the most effective in penetrating strongly cohesive oxides, the serious disadvantage of this group is the corrosive residue which remains after soldering.

Although the residue may be neutralised there is always a possibility that a small percentage may become trapped and escape neutralisation. Also it is an added expense to carry out this process, which may not be acceptable for electronic assemblies.

These disadvantages do not apply in the case of a gas, but this method is rather difficult to use because:

(*a*) a large, and therefore expensive, gas chamber would be required, capable of holding a mechanical soldering system,
(*b*) hydrogen is generally used as the reducing agent; this gas is expensive and can be dangerous, especially at elevated temperatures.

For these reasons inorganic fluxes are rarely used in the electronics industry.

Organic Fluxes (Corrosive)

This group, termed intermediate corrosive fluxes, are weaker in action than the inorganic group. However, the residue may still be corrosive, but to a lesser degree than the first group. Their use is limited to lower temperatures than the highly corrosive variety, because they are not as stable at higher temperatures as the first group of fluxes.

TABLE 1

Comparison of Flux Materials in order of decreasing activity

Classification and type	Typical fluxes	Vehicle	Use for these joints	Approx. max. working temperature	Tarnish removal	Corrosivity	Postsolder cleaning methods
INORGANIC Acids	Hydrochloric, hydrobromic, phosphoric	Water, petroleum jelly	Structural	600°C	Very good	High	Hot water rinse and neutralise
Metal salts	Zinc chloride, tin chloride	Water, petroleum jelly, polyethylene glycol	Structural	600°C	Very good	High	Hot water rinse and neutralise
Non-metal salts	Ammonium chloride, hydrazine hydrochloride	Water, urea	Structural	350°C	Good	Moderate	Hot water rinse and neutralise
Gases	Hydrogen	None	Electrical	550°C (Spontaneous ign. temp.)	Very good	None	Not needed
ORGANIC (non-rosin) Acids	Lactic, oleic, stearic, glutamic, phthalic.	Water, organic solvents, petroleum jelly, polyethylene glycol	Structural, Electrical	400°C	Fairly good	Moderate	Hot water rinse and neutralise; organic solvents*
Halogenated	Aniline hydrochloride, glutamic acid hydrochloride, bromide derivatives of palmitic acid	Water, organic solvents, petroleum jelly, polyethylene glycol	Structural, Electrical	400°C	Fairly good	Moderate	Hot water rinse and neutralise; organic solvents*
Amides and amines	Urea, ethylene diamine	Water, organic solvents, petroleum jelly, polyethylene glycol	Structural, Electrical	400°C	Fair	Moderate	Hot water rinse and neutralise; organic solvents*
ORGANIC (rosin-base) Activated	Rosin + amine hydrochloride	Organic solvents	Electrical	350°C	Fair	None normally	Not normally needed; organic solvents if required
Non-activated	Pure rosin	Organic solvents	Electrical	350°C	Poor	None	

* For more thorough cleaning, follow by demineralised or distilled water wash.

The information in the above table is taken mainly from H. H. Manko, 'How to Choose the Right Solder Flux' (Ref. 3). Further information has been added to this summary of flux materials. This latter information is based on material gathered from sources in Mullard and associated companies, and in some instances differs from the table given in the reference above.

In many electronic fields, these fluxes are not acceptable because of the danger of corrosion and the necessity for neutralisation.

Organic Fluxes, Rosin Based (Non-corrosive)

This group of fluxes may be divided into two sections; non-activated and activated fluxes. Both consist of a grade of natural abietic acid which, though usually described as water-white, are pale straw in colour. Although not as effective as the first two groups at removing oxides, this type of flux has gained almost universal acceptance in the electronics industry since the flux residue after soldering is non-corrosive, hard, non-hygroscopic and non-conducting. There is, therefore, no need for flux residue removal which would otherwise present a very difficult problem in a complex assembly.

Unfortunately rosin (natural abietic acid) has only a slow and mild action and, for most practical purposes, has to be boosted by the addition of an activator. This is usually an amine hydrochloride; the quantity added must not exceed 0·5% chloride for a liquid flux. For a cored solder, up to 0·7% activator is permissible. Provided a correct choice of activator is made, the flux residue has properties virtually identical with those of pure rosin.

As a general rule, the rosin-based fluxes are the only type sufficiently non-corrosive for electrical use. Since they have only a limited oxide-removing capacity, however, components intended to be soldered must not have more than a light oxide deposit on the surfaces to be soldered. Their relatively low upper temperature limit is not usually a serious disadvantage since other factors (such as the temperature stability of transistors) often impose an even lower limit.

SOLDER ALLOYS

There are many different solder alloys in use, most of them being for rather specialised applications, but the alloys most used for electronic purposes are composed simply of a mixture of tin and lead. Whatever the alloy, there are three important temperatures related to its use. These are:

(i) The solidus, the highest temperature at which the alloy is completely solid

(ii) The liquidus, the lowest temperature at which the alloy is completely liquid

(iii) The superheat, the temperature above the liquidus at which the solder will run freely on to the joint.

The solidus and liquidus temperatures depend only on the composition of the solder alloy, whereas the superheat depends also on the heat capacity of the joint members and of the heat source (whether soldering iron or solder bath).

The region between the solidus and liquidus is referred to as the freezing range, melting range or pasty range. Between these temperatures the alloy is a pasty mixture of liquid and solid. Fig. 2 shows how these temperatures are related to the proportion of tin (or lead) in the alloy. It may be seen that an alloy containing 63% tin, 37% lead has no pasty range, the liquidus and solidus temperatures being identical. Such a composition is referred to as a eutectic mixture; like pure tin or lead, it has only a single melting point. All other compositions have a liquidus temperature higher than the melting point of the eutectic; 50% tin, 50% lead, for instance, has a liquidus temperature of 214°C and a solidus temperature of 183°C.

For general-purpose electronic soldering the alloy most commonly used contains 60% tin, 40% lead. Where the joint may have to withstand a high working temperature as, for example, in the connection of mains droppers in domestic receivers, a solder of higher melting range must be used. From Fig. 2 it would seem logical to use the solder with the highest melting temperature, namely pure lead. However, a disadvantage with the use of pure lead is the inherent poor flow characteristics. By the addition of small quantities of tin and silver this difficulty is overcome, at the expense of a small reduction in melting temperature.

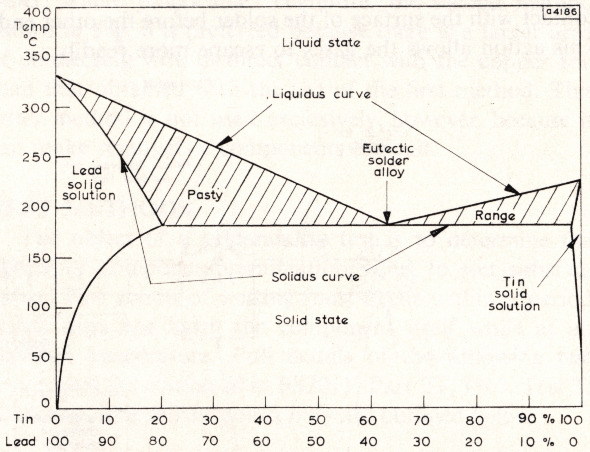

Fig. 2—Solder alloy composition temperatures

MECHANICAL SOLDERING TECHNIQUES

Mechanical soldering techniques may be readily applied to printed circuit panels where the surface to be soldered is flat. A simple method of soldering, which is widely practised, is to dip the panel to a predetermined level in a solder bath after flux has been applied.

The molten solder is maintained at the correct soldering temperature, 235°C±2degC, by a thermostat, situated in the bath.

The panel, which is at room temperature, is dipped into the molten solder mass where thermal conduction causes it to be raised to the solder temperature, and wetting takes place. Thermal inertia of the mass of solder in the pot compensates for any small losses resulting from heating the joint areas. Thus the heated solder performs two functions; it heats the work surfaces to the soldering temperature and wets the joint.

Oxides or dross form rapidly on the surface of a molten solder bath and these contaminations, which prevent wetting, must be removed before any component is dipped. This difficulty may be overcome by the use of wiper blades with which the solder surface may be swept before each dipping.

Flux is applied to work surfaces before dipping as a protection and an aid to wetting, as mentioned earlier. However, it is imperative that the solvent used in the flux evaporates as wetting occurs, since the flux is displaced by molten solder at the wetting temperature. It is possible that some flux solvent can become trapped beneath a layer of molten solder and, in escaping in vapour form, cause the solder to be forced away from the work surface; this bubbling has a deleterious effect on the solder joint.

To facilitate volatilisation of the solvent, it is desirable to lower a printed panel so that one end of it comes into contact with the surface of the solder before the other end. This action allows the gases to escape more readily.

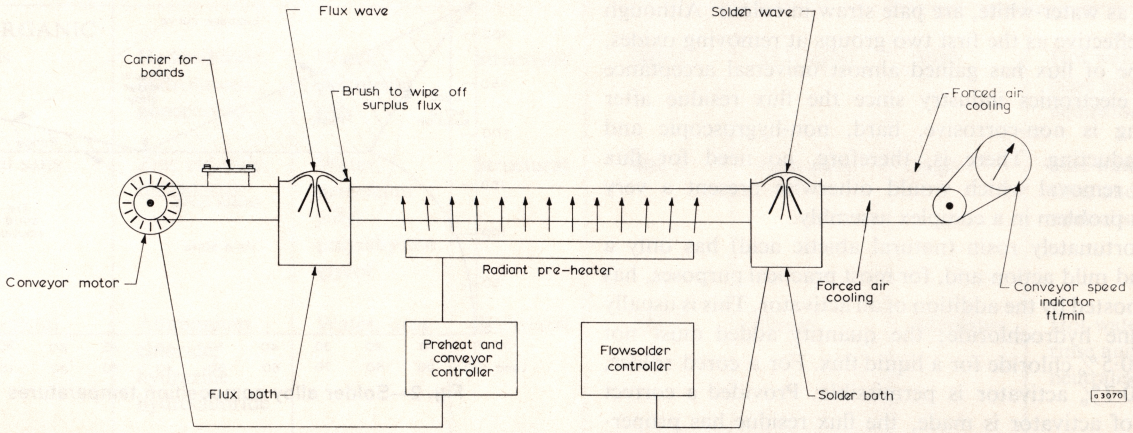

Fig. 3—Flowsolder operation cycle

Whereas an angle on insertion of 2° to 3° from the horizontal may suffice to allow the solvent gases to escape, an angle of approximately 30° on withdrawing the panel may be found advantageous. The steeper angle assists in preventing contacts becoming bridged with surplus solder by allowing any surplus to drain back into the bath.

In recent years dip soldering has been improved in several ways. One example is the method of using a mechanical lever to dip a printed panel to the correct depth. The lever slowly places the panel down into the solder bath at an angle which allows the flux solvent to escape. The panel is then removed and mechanically vibrated, which shakes surplus solder back into the bath.

Another improvement to dip soldering which obviates the dangers of bridged contacts and trapped solvent is the use of a rake method. A bar with a series of adjustable prongs placed at right angles to it – with the appearance of a rake – is immersed in a solder bath. At the end of each prong is a small cup. The cups are aligned to coincide with joints on a panel situated above the bath, so that when the cups are swung up to the panel the flux and then the solder are placed precisely on the joint areas only.

This type of soldering is especially suitable with, for example, printed coils where small joints are made and it is imperative to avoid short-circuiting any part of the coil.

Another mechanical soldering method which has been used with success is the flow or wave solder technique. Because flow soldering is relatively new in comparison with dipping, a detailed account of this technique is given in the next section.

Flow Solder Equipment

Fig. 3 gives a pictorial impression of the flow solder cycle used on Fry's Flowsolder Machine. The principle of operation is that the underside of the printed board cuts the crest of a molten solder wave as it travels along on a conveyor. The wave is produced by an immersed impellor pump forcing molten solder up through a constricted slot, formed by two angled converging plates. This method results in a solder wave which has a smooth level crest and is always clean and free from oxides and dross (Fig. 4). Freedom from contamination is achieved because the solder which is forced into the constriction to form the wave is taken from below the surface of the bath. Also, because the solder is being pumped in a continuous stream from the bottom of the bath, it is at a constant temperature. The height of the solder wave may be adjusted by altering the speed of the impellor pump.

The printed panel passes in succession over a flux wave, a pre-heating area, and the solder wave. Flux is applied to the underside of the printed panel in a similar manner to that used with the molten solder wave, and the panel then traverses a preheating area where the flux is partly dried. The preheating removes some of the solvent from the flux and helps to prevent excess solvent evaporating in the solder bath. Heating the printed panel also assists in preventing the molten solder becoming chilled.

Fig. 4—Contamination-free solder wave

In some systems the flow solder cycle is completed by using forced draught cooling on both sides of the panel as it emerges from the solder wave. This procedure reduces the highest temperature reached by the panel, as indicated by the dotted line in Fig. 5.

Fig. 5 shows a typical curve of a semiconductor junction temperature versus the time taken for a printed board flow-solder cycle. The rise in junction temperature is determined mainly by the preheat temperature and the time taken to traverse the preheat area, since rapid heating and cooling of the junction occurs when the device leads are actually in contact with the molten solder. The thermal 'pulse' is the result of heat flowing up the lead to which the semiconductor crystal is attached. Isolating this lead with a thermal shunt during the flow solder cycle would remove the heat pulse almost completely.

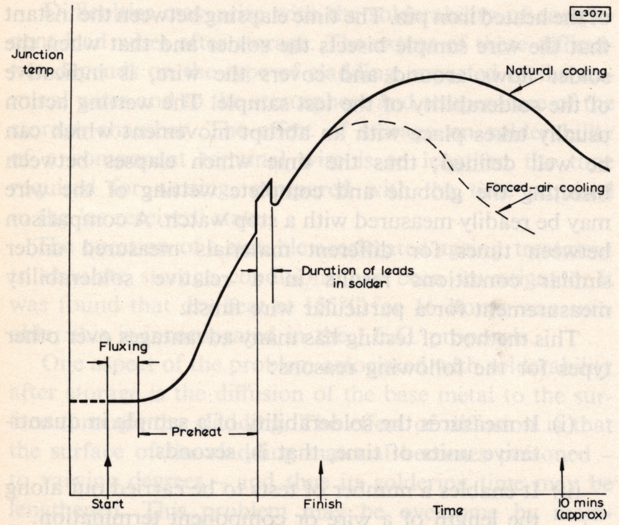

Fig. 5—Junction temperature of a semiconductor device mounted on a panel, versus time taken for flow solder cycle

The second, more gradual, peak is caused by thermal conduction through the printed circuit board and the body of the semiconductor. Some American government authorities specify an oil film on the solder to reduce surface tension and allow a lower soldering temperature without loss of effectiveness. This technique helps to protect temperature-sensitive components.

COMPONENT PRECAUTIONS

The majority of components used in any form of electronic circuitry are susceptible to thermal and/or mechanical damage during assembly processes. Limitations or tolerance for soldering, if any, may vary from one type of component to another according to its susceptibility to damage; this information is given in the published data for the device concerned. Therefore, it is essential that data for each individual component is consulted to ensure that its specified tolerances are not exceeded by the assembly method used. Where a number of components are being used in a printed circuit panel, the device which is the most sensitive to thermal damage governs the thermal restrictions which apply for a single mechanical soldering process.

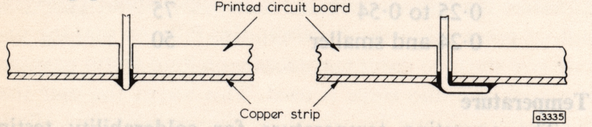

Fig. 6—Two methods of mounting component leads on a printed panel

Fig. 6 illustrates two methods of inserting component lead wires through an appropriate aperture in a printed panel, before soldering takes place.

To obtain good mechanical joint strength, it is desirable to have large surface areas in direct contact with each other, where practicable. Therefore, the second method shown in Fig. 6 is preferred because there is a larger area of connecting wire in direct contact with the copper foil than that obtained with the use of the first method. This latter method is not used exclusively, however, because it can make removal of components difficult.

TEST METHODS

The object of a solderability test is to determine the ability of component termination wires to wet satisfactorily. This action of wetting must occur within a period which does not harm the component itself while at an elevated temperature. Full details of the following test procedures are contained in BS2011: Part 2T, 1963, Test T, Soldering. The salient points only are outlined below.

Solder Bit Test

A soldering iron is used where the bit size is a specified diameter; for example, 8mm. An exposed bit length of 32mm which is reduced to a wedge shape over a length of approximately 10mm is defined.

The surface of the bit must be smooth and properly tinned; 60% tin, 40% lead solder with a non-activated wood rosin core is used. The soldering temperature is between 300 and 350°C.

The test method is to apply the iron and solder to a test wire. The wire must be fully tinned to within 6mm of the fixing nut of the body or half the length of the termination tags, at the end of 2s.

Solder Bath Test

The solder bath is fitted with a thermostat and has a volume which is sufficient to ensure that the temperature of the solder remains constant when the specimen is introduced into it. The exposed surface area of the solder must be approximately 5cm^2 to reduce the effect of direct radiation on the component under test.

The surface of the bath must be kept clean, and immediately before the immersion of a test sample a piece of solder is dropped into the middle of the bath. The solder is specified as 12mm length, 1·6mm diameter, of 60% tin, 40% lead alloy with a non-activated wood rosin core. No other flux is required.

There are two temperatures specified; 230°C±10degC for components intended for printed wiring applications; 270°C±10degC for other components.

The procedure is for the component wire to be immersed for 2s±0·5s to a point 6mm from the device, as soon as the added piece of solder has melted.

The wire is then examined for good tinning, as evidenced by free flowing of the solder.

Solder Globule Test

The principle of this test is to insert a wire into a preweighed solder globule which is maintained at a temperature of 235°C±2degC, when wetting of the test wire by the solder globule may be observed and timed (Fig. 7).

Modifications have been carried out to the original test equipment, mainly in respect of the method of heating the solder globule, by an E.E.A. working group. Closely comparative results have been achieved with the use of the modified equipment over a large number of test samples.

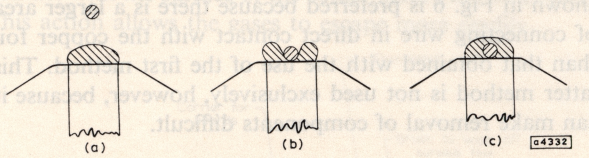

Fig. 7—Solder globule test principle

The equipment is intended for testing the solderability of round terminal wires up to 1·2mm in diameter, and it is therefore especially suitable for the electronics industry. The modified test equipment is now included in the latest draft revision of BS2011 Part 2T.

This test equipment (Fig. 8) consists of an aluminium block into which is inserted a spectrographically pure iron pin of circular cross section, flush with the top surface of the block. The block is heated by a thermostatically controlled element which is introduced into it. The iron pin is tinned and a preweighed solder pellet is placed on the pin, heated and fluxed.

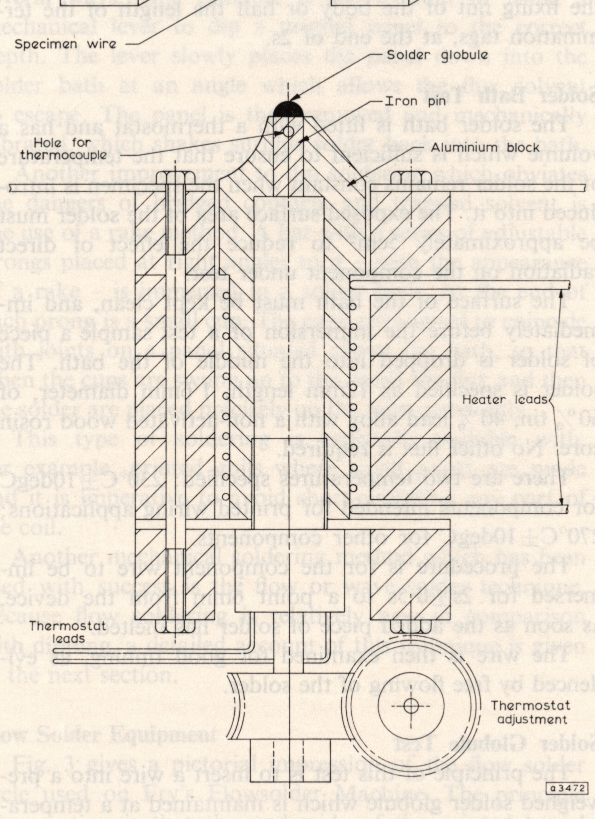

Fig.8 —Solderability test equipment

The pellet melts to a globule and remains on the pin, which acts as a locating base for it. Thus good thermal conductivity is accomplished because the solder globule is located precisely and securely on the pin. The test wire, which is held substantially straight in a clamp over the molten solder, is lowered to meet the solder.

The wire bisects the globule and touches the surface of the heated iron pin. The time elapsing between the instant that the wire sample bisects the solder and that when the solder flows around and covers the wire, is indicative of the solderability of the test sample. The wetting action usually takes place with an abrupt movement which can be well defined; thus the time which elapses between bisecting the globule and complete wetting of the wire may be readily measured with a stop watch. A comparison between times for different materials measured under similar conditions results in a relative solderability measurement for a particular wire finish.

This method of testing has many advantages over other types for the following reasons:

(i) It measures the solderability of a sample in quantitative units of time, that is, seconds.
(ii) It enables a number of tests to be carried out along the length of a wire or component termination.
(iii) It is discriminating in its action.
(iv) It enables solder tests to be carried out without severance of the connection wire.

During a trial period of this equipment a great many wire samples and component terminations having a diversity of finishes were tested. The results on the latest equipment for tests held at different establishments were compatible for similar wire finishes. Consistent results have also been obtained by the International Electrotechnical Commission working groups in the Netherlands and Sweden.

TEST CONDITIONS

After a considerable amount of research and discussion a working party of the I.E.C. has agreed to the following proposals for solderability tests.

Solder

The solder to be used in these tests should be 60% tin, 40% lead solder in pellet form. The size of pellet is determined by the wire diameter under test, and the following relationship applied:

Wire Diameter (mm)	Pellet Size (mg)
0·75 to 1·2	200
0·55 to 0·74	125
0·25 to 0·54	75
0·24 and smaller	50

Temperature

The operating temperature for solderability testing should be 235°C±2degC. These limits apply for tests on components intended for both conventional and printed wiring circuits.

Flux

The flux should be 25% by weight of rosin grade 'WW' in 75% by weight of 99% isopropyl alcohol to BS1595.

STORAGE OF COMPONENTS

Difficulties may arise with the solderability of component lead wires after storage. The extent of these difficulties depends on the type of cladding imparted to the terminal wires and to the atmosphere and temperature of the storage chamber. The effect of storage on solderability of a component terminal wire is to lengthen the time required for wetting, compared with the time required in the 'as received' state.

The question of a suitable accelerated ageing treatment to simulate storage conditions has been investigated. It was found that dry heat at 155°C for 16 hours was suitable; this is incorporated in the I.E.C. proposals.

One aspect of the problem associated with solderability after storage is the diffusion of the base metal to the surface through the cladding. The effect of diffusion is that the surface of the cladding material becomes poisoned – to varying degrees – and thus its soldering time may be lengthened. This problem may be overcome by maintaining a minimum thickness of surface material, when diffusion into it is more difficult and unlikely.

At the present time it is considered as a guide by the I.E.C. that sample testing of terminal lead out wires up to 1·2mm diameter should meet the following limits for solderability: a maximum soldering time of 4s for any component termination wire when examined to DEF131A, inspection level S4, to an AQL of 4%.

SOURCES

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Voltage Stabiliser for Mains/Battery Television Receivers

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The requirements for a voltage stabiliser for use in a television receiver are discussed as an introduction to a complete circuit design. The circuit provides an output of 11V at 2·5A maximum and is safe for any load condition, the current being limited by the circuit. The output resistance is 0·3Ω, and the hum level is about 150mV pk-pk.

INTRODUCTION

Stabilisers are at present used in mains/battery transistor television receivers for two major reasons. The first object of the stabiliser is to keep the picture height and width constant on battery operation as the battery discharges. Secondly, since a stabiliser supplies a very constant output potential, the line output transistor can be used closer to its maximum ratings. There are also additional advantages in that the mains power supply need not have either a good regulation or a ripple-free output when a stabiliser is used between it and the receiver. The principle of operation is discussed as an introduction to the circuit design.

BASIC STABILISER CIRCUIT

The basic circuit arrangement for a voltage stabiliser is shown in Fig. 1, and may be considered in two separate sections:

(*a*) The differential amplifier which supplies a control current as a function of the output voltage, and

(*b*) the series regulator whose current, and hence impedance, is controlled by the current drive from the differential amplifier.

Differential Amplifier

Output potential variations are applied to the differential amplifier together with a d.c. reference potential. In the circuit shown in Fig. 1 the output variations are applied to the emitter via the v.d.r. across which the reference potential is developed, the base being returned to a potential divider across the stabiliser output.

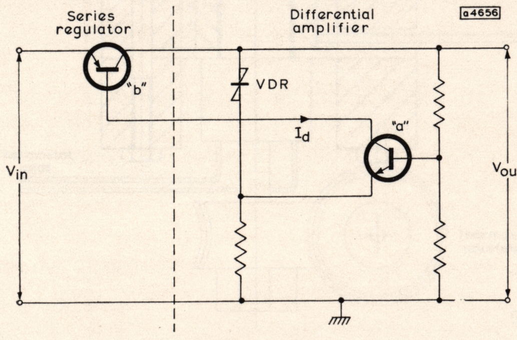

Fig. 1—Basic stabiliser circuit

If we now consider a small drop in the output potential of the stabiliser, an increase of the applied V_{be} to the differential amplifier will result (typically about half the variation in V_{out}), which makes the differential amplifier pass a higher collector current. This collector current is fed to the series regulator transistor to control the stabiliser output, and is referred to as the drive current I_d.

Series Regulator

The series regulator transistor controls the current through the stabiliser and hence determines the output potential. The series regulator transistor is controlled by the drive current I_d from the differential amplifier. Because of the large swing in base current required by the series regulator transistor and the small values of I_d available from the differential amplifier, an emitter follower (TR_2 in Fig. 2) is used to drive the series transistor. To reduce the maximum dissipation in the series transistor and the emitter follower at high input potentials, a shunt resistor may be used across the series transistor (TR_1 in Fig. 2).

DESIGN CONSIDERATIONS AND CIRCUIT DESCRIPTION

Battery Operation

The potential of a car battery is about 15V when fully charged. It reduces rapidly to some 13V with load and then spends most of its life between 13V and 12V, after which it again reduces rapidly until discharged. The battery potential should be above 11·5V for about 70% of its life from full charge. For optimum battery usage the widest part of the battery life characteristic must be used, hence the lowest terminal voltage consistent with good receiver performance is chosen. By using a series regulator circuit, as shown in Fig. 2, a bottoming potential of less than 0·5V (typically 0·25V) can be achieved for the series transistor. The output potential of the stabiliser is therefore taken as 11V.

When the battery potential falls below about 11·5V the output potential of the stabiliser also falls, so that the picture size starts to reduce, especially on loud audio passages if a class B audio stage is used. This warns the user that the battery should be recharged, but nevertheless allows further viewing of the receiver.

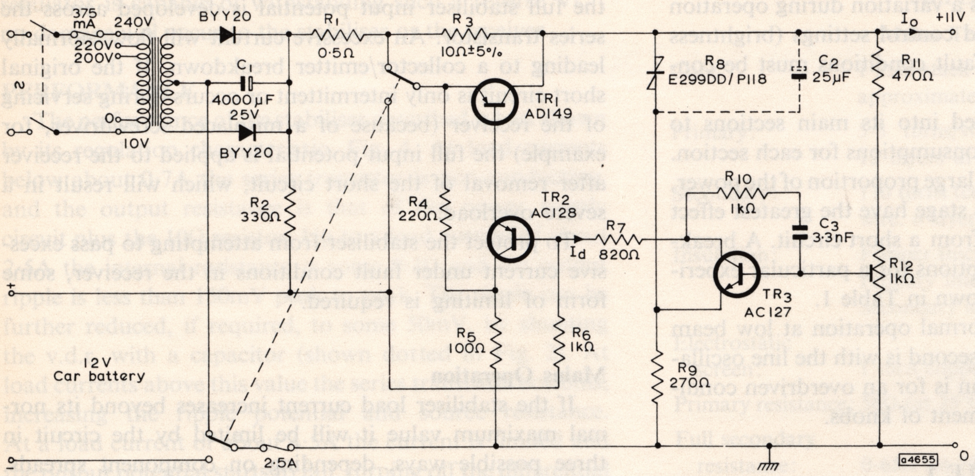

Fig. 2—Practical stabiliser circuit for mains or battery operation. The transformer specification is given at the end of the article

The maximum dissipation in the series transistor occurs with a fully charged battery at maximum load current. Assuming this to be 2A, the dissipation in the series transistor is some 8W.

Mains Operation

The series regulator transistor will have maximum dissipation at high mains input potentials and maximum load current. Since the output potential from the mains supply unit has a ripple potential which is effectively smoothed by the regulator circuit, the effective input potential, and hence the dissipation in the series transistor, is higher than on battery operation. The maximum dissipation in the series transistor at 2A load current is approximately 12W. To reduce the power dissipation on mains operation a resistor can be used in parallel with the series transistor.

In a typical design, such as the one described, the output from the power supply unit at the input to the series transistor is 21·5V (nominal mains, no load current) falling to 13·5V at maximum current (2A) and low mains inputs (−10%) (see Fig. 3). The ripple potential is 3V peak-to-trough with the trough occurring some 2V below the mean value, so that the minimum input to the regulator transistor (TR_1 emitter) is 11·5V, hence the transistor is never bottomed by the ripple potential.

The shunt resistor is chosen to just pass the minimum load current at high mains inputs. In the design described the minimum load current is taken as 1A and the maximum input potential at this current is almost 21V (see Fig. 3). Hence the value of the required shunt resistor is 10Ω. The maximum dissipation in this resistor is 3·6W, hence the dissipation in the series transistor is 8·4W.

The series regulator transistor is an AD149 used in conjunction with a heatsink. The required overall thermal resistance θ_{j-a}, for a maximum ambient temperature of 50°C and a maximum permitted junction temperature of 100°C, is therefore 50degC/8·4W = 5·95degC/W. Since the thermal resistance of the transistor plus mica washer is 2·5degC/W, the heatsink must have a thermal resistance not greater than 3·45degC/W. In practice a $6\times6\times\frac{1}{16}$in sheet of blackened aluminium was used.

Differential Amplifier

The v.d.r. is chosen to operate at approximately half the stabiliser output potential at a convenient current. The current must be high enough to operate the v.d.r. at a low slope resistance and be capable of supplying the maximum differential amplifier emitter current. Variation of the output potential will occur due to heating of the v.d.r., hence the maximum v.d.r. current is limited by this consideration.

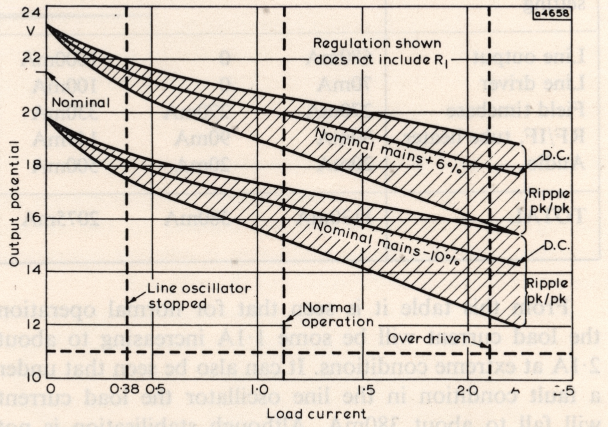

Fig. 3—Power supply unit regulation (240V tap). The regulation shown does not include R_1

In the design described a Mullard v.d.r. type E299DD/P118 is operated at 20mA, and it has a slope resistance of 90Ω.

To avoid the base current of the differential amplifier significantly influencing the base potential, a low-resistance potential divider is required. A range of adjustment is provided by means of R_{12} to allow for component tolerances.

The resistors constituting the load of the differential amplifier must have as high a value as possible from gain considerations, but the maximum value is limited by leakage current in the emitter follower at high temperatures. The feedback circuit, R_{10} and C_3, reduces the gain at high frequencies and prevents oscillation due to phase shift in the transistors.

The power dissipation in the emitter follower transistor is reduced by feeding its collector from a potential divider R_4 and R_5.

REQUIRED RANGE OF LOAD CURRENTS

A television receiver presents a varying load to the power supply. Firstly there is a variation during operation depending on the signals and control settings (brightness and volume); furthermore, fault conditions must be considered.

A receiver may be divided into its main sections to obtain the range of current consumptions for each section. The line timebase consumes a large proportion of the power, hence fault conditions in this stage have the greatest effect on the power supply apart from a short circuit. A breakdown of the current consumptions for a particular experimental 11-inch receiver is shown in Table 1.

The first column is for normal operation at low beam current and low volume, the second is with the line oscillator stopped. The third column is for an overdriven condition; that is, with maladjustment of knobs.

TABLE 1
Current Consumption in Television Receiver

Condition	Normal operation	Line oscillator stopped	Overdriven condition
Beam current setting	50μA	0	150μA
Line output	680mA	0	1000mA
Line driver	70mA	0	100mA
Field timebase	270mA	270mA	330mA
RF/IF, tube heater	90mA	90mA	145mA
Audio	20mA	20mA	500mA
TOTAL	1130mA	380mA	2075mA

From this table it is seen that for normal operation the load current will be some 1·1A increasing to about 2·1A at extreme conditions. It can also be seen that under a fault condition in the line oscillator the load current will fall to about 380mA. Although stabilisation is not necessary under these conditions, the degree of output potential rise must be limited to protect the receiver. The voltage/current characteristic for mains operation is shown in Fig. 4.

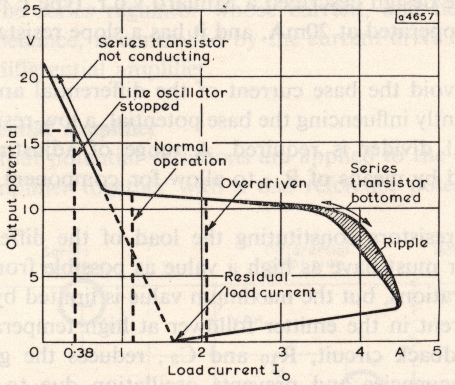

Fig. 4—Stabiliser output characteristic (nominal mains)

FAULT PROTECTION

In the event of a short circuit occurring in the receiver, the full stabiliser input potential is developed across the series transistor. An excessive current will flow, normally leading to a collector/emitter breakdown. If the original short circuit is only intermittent or occurs during servicing of the receiver (because of a misplaced screwdriver, for example) the full input potential is applied to the receiver after removal of the short circuit, which will result in a severe overload.

To protect the stabiliser from attempting to pass excessive current under fault conditions in the receiver, some form of limiting is required.

Mains Operation

If the stabiliser load current increases beyond its normal maximum value it will be limited by the circuit in three possible ways, depending on component spreads. Firstly, the series regulator transistor will bottom, due to the power supply source impedance. Secondly, the emitter follower tends to bottom if the h_{fe} of the series transistor is low. Thirdly, as the output potential falls in the particular circuit configuration chosen, the differential amplifier will be turned off, causing a still further fall in the output potential.

In the circuit configuration shown (Fig. 2) the falling output potential applied via the v.d.r. normally tends to drive the differential amplifier harder into conduction. This current is limited by R_9, and the potential drop across R_9 is in effect negative feedback. In the event of the h.t. line potential falling significantly, the potential across the v.d.r. also falls, increasing its resistance, and the feedback effect of R_9 is correspondingly increased. This effect, coupled with the fall in base potential applied by the potential divider R_{11} and R_{12}, turns the differential amplifier, and hence the series regulator transistor, off.

When a resistor is connected in parallel with the series regulator transistor the residual load current will be determined by its value and the regulation of the power supply. With the circuit shown in Fig. 2 the residual load current is about 1·5A as shown in the complete output characteristic of Fig. 4.

Battery Operation

Fault protection on battery operation is more difficult than with mains operation since the battery source impedance is much lower than that of the mains power unit. As no parallel resistor is normally necessary across the series regulator on battery operation, the short circuit current will be practically zero. However, a partial short circuit can result in a much higher series regulator current, but the series regulator dissipation will depend to a large extent on the battery potential. In these circumstances, the main protection will be provided by the input fuse. However, fault protection on battery operation is less important than mains operation since most of a receiver's usage will be with a mains input, and servicing will normally be carried out with a mains supply.

In the event of reversed input voltage polarity to the stabiliser due to incorrect connection of the battery, practically all of the input potential is dropped across the series regulator emitter/base junction. As the series regulator is capable of withstanding such a reverse V_{be}, no damage will occur to the stabiliser or the receiver.

PERFORMANCE

The performance of the stabiliser described is best shown by its regulation characteristic, Fig. 4. At load currents below about 0·7A the series transistor is non-conducting, and the output resistance is that of the power supply circuit plus the 10Ω resistor. Up to a load current of about 2·5A the internal resistance is some 0·3Ω and the output ripple is less than 150mV peak to peak. The ripple can be further reduced, if required, to some 50mV, by shunting the v.d.r. with a capacitor (shown dotted in Fig. 2). At load currents above this value the series transistor bottoms, increasing the ripple potential and source resistance. At a load current of about 4·5A the current is limited and the output potential starts to fall, turning off the differential amplifier. The output resistance is then once again that of the power unit plus the 10Ω resistor. The shape of the overload characteristic ensures that the regulator is completely safe with a short circuit load.

APPENDIX

Mains Transformer

Core:	1·25in stack of 29A Silcor 107 laminations approximate weight = 2lb. M.E.A. Ltd.
Primary:	5·3 turns/volt of 33 s.w.g. en.cu. wire. 12 layers at 111 turns/layer.
Secondary:	170 turns (centre tapped) of 20 s.w.g. en.cu. wire. 6 layers at 29 turns/layer.
Insulation:	Former thickness = 0·025in. Primary interleaving thickness = 0·002in. Secondary interleaving thickness = 0·004in.
Electrostatic screen:	1 turn copper foil 0·02in.
Primary resistance:	56Ω (at 25°C)
Full secondary resistance:	0·65Ω (at 25°C)
Magnetising current:	46mA (240V tap + 6% mains).
Transformer efficiency:	84% at 37W output (90°C).

Industrial R.F. Oscillator for 150Mc/s

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The YD1150 and YD1160 series of industrial triodes offer advanced design features. It is shown how these can be utilised to construct an industrial r.f. oscillator at 150Mc/s. The preferred form of circuit, of which constructional details are given, can be used for either the YD1150 series to give 3·2kW load power or the YD1160 series to give 7·0kW load power. An overall efficiency up to 70% is obtained.

INTRODUCTION

A range of power triodes with ceramic envelopes, suitable for use in industrial r.f. generators, is becoming available. Of this range, the YD1150 series and the YD1160 series are both capable of operation at full input ratings up to 160Mc/s. A circuit for these types at such a frequency was accordingly constructed.

VALVE AND CIRCUIT CONSIDERATIONS

In the design of v.h.f. generators the importance of joining the valve in the correct manner to the circuit increases with increasing frequency if efficient operation free from spurious responses is to be achieved.

In a basic oscillator where no upward transformation is intended, the maximum r.f. potentials occur at the thermionically active electrode faces of the valve, which are also the loci of the nominal inter-electrode capacitances. As frequency increases, these capacitances become more important and contribute to the total capacitive component of the oscillatory circuit. All external reactances must thus be connected in the shortest possible way, that is with the minimum of intervening inductive reactances. Such a construction can obviously only be realised if the lead inductances from the active valve surfaces to the valve exterior are already suitably short and substantial.

The engineering problems normally resulting from the use of conventional valves are greatly eased by the ceramic-coaxial construction of the new series of industrial valves. Connections of high current carrying capability and little stray reactance may readily be made to the coaxially terminated electrodes by means of finger-strip rings or directly attached plates. One or more of these plates may then be conveniently used as carriers for the feedback system. Their mechanical and hence electrical closeness to the valve electrode system will help to maintain the correct phase relationship for efficient operation.

Tank Circuit

The frequency controlling section is the tank circuit between grid and anode in the form of a capacitively loaded line. The loading capacitance, as is normal practice with v.h.f. oscillators, is formed by the inherent c_{a-g} of the valve alone.

So that for the YD1150			and for the YD1160
c_{a-g}	=	14pF	19pF
P_{in}	=	5kV×1A	5kV×2A
f_0	=	150Mc/s	150Mc/s
Q_L	=	39	25

Both Q_L values are acceptable for oscillatory operation, and the design of the line is based on the YD1160, the physically larger valve.

If a line circuit of minimum circuit loss is to be designed, its characteristic impedance should be near the theoretical optimum of $Z_0 = 77\Omega$. The air cooler of the YD1160 having a diameter of 122mm is thus suitable to form a line inner of a circuit that is commensurate with the power level of the valve, and will for the given impedance require a line outer of 3·6 times the valve diameter or 440mm.

Whilst these considerations apply to lines of circular cross-section, little is changed if, for ease of construction, the outer is given a square cross-section with heavily radiused edges. The side length for such a box, as used in the present example, is then 430mm. The line length or height of this box is calculated by standard methods and for the given frequency is (with $c_{a-g} = 19$pF and additional stray capacitance of 3pF from the top end of the cooler) 160mm. To bring the end plate of the line box to the level of the grid ring on the valve this dimension was reduced to 150mm. The resulting change in frequency is less than 3%.

From the sectional drawing in Fig. 1 it can be seen that the above derived line dimensions, giving an r.f. current path of square cross-section, may be regarded as a lumped circuit with a favourable ratio of inductive path length to enclosed area, approaching the ideal toroidal cavity. Circuit losses are therefore expected to be small.

Full dimensional details are given in Fig. 1, whilst Figs. 2 and 3 give a general view and internal detail of the tank circuit. P.T.F.E. sandwich capacitors extending over the large faces of this line box serve as anode and grid blocking capacitors.

Feedback System

With the given tank circuit, oscillatory operation may be achieved, either in a Colpitts system or by the addition of a grid-filament line, converting the circuit into a tuned-grid tuned-anode earthed-grid system. Both systems require additional C_{a-f} for correct functioning.

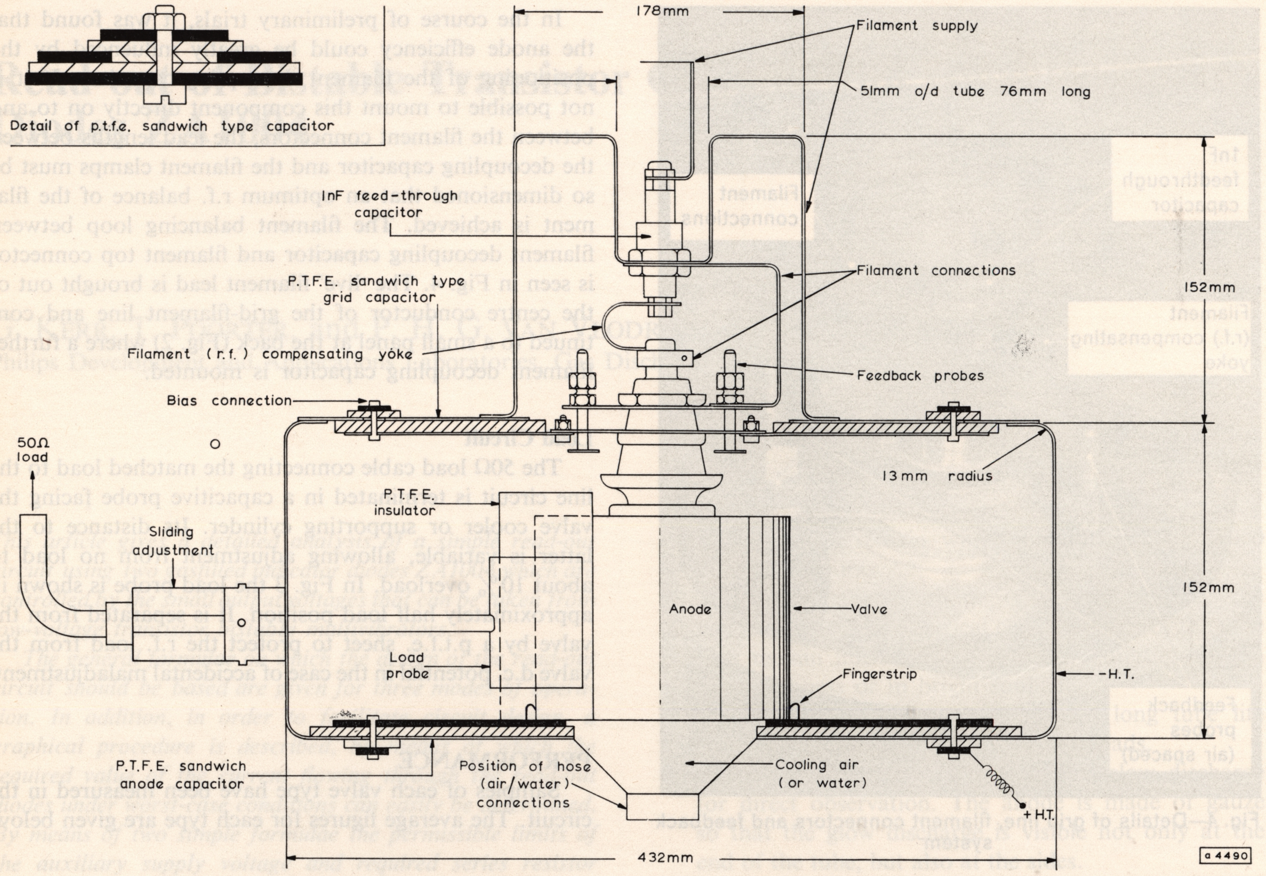

Fig. 1—150 to 165Mc/s circuit for YD1150—YD1160 series

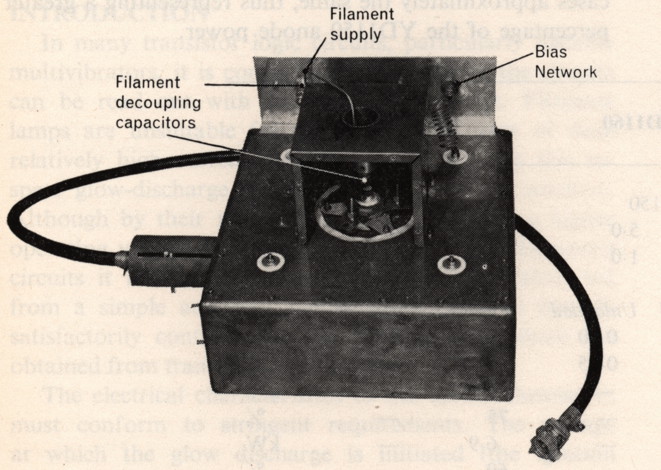

Fig. 2—150 to 165Mc/s cavity circuit complete

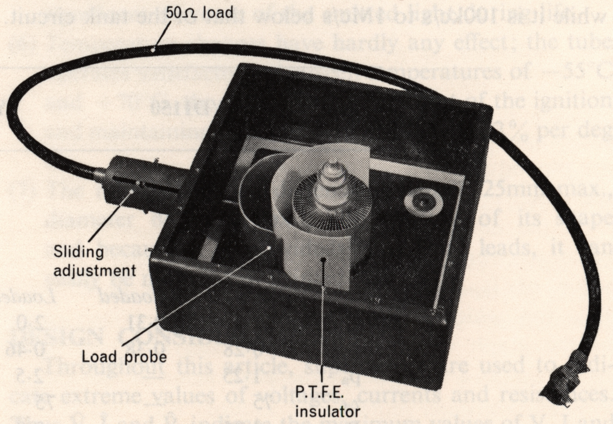

Fig. 3—Cavity with top removed

The originally proposed feedback system consisting of three plungers extending from the larger filament connector, and mating with three corresponding p.t.f.e. lined cups at anode level, had in earlier experiments given rise to ionisation, once the circuit-valve assembly had reached operating temperature. Air spaced probes of equivalent capacitive value were therefore used. These are shown attached to the valve in general view in Fig. 2 and in close up in Fig. 4, in which two of the air spaced probes can be clearly seen.

Filament Circuit

With the proposed tank circuit construction the filament connectors will be above the level of the grid connector and therefore outside the cavity. The feedback probes extending from the lower filament connector must thus re-enter the tank cavity through apertures that are small compared with the wavelength. Therefore to prevent potential changes (incorrect feedback phasing) along the length of these probes, the filament circuit must be matched and screened. This is conveniently done with the help of a $\frac{1}{4}\lambda$ resonant line between grid and filament and capacitively loaded by the c_{g-f} internal valve capacitance. The con-

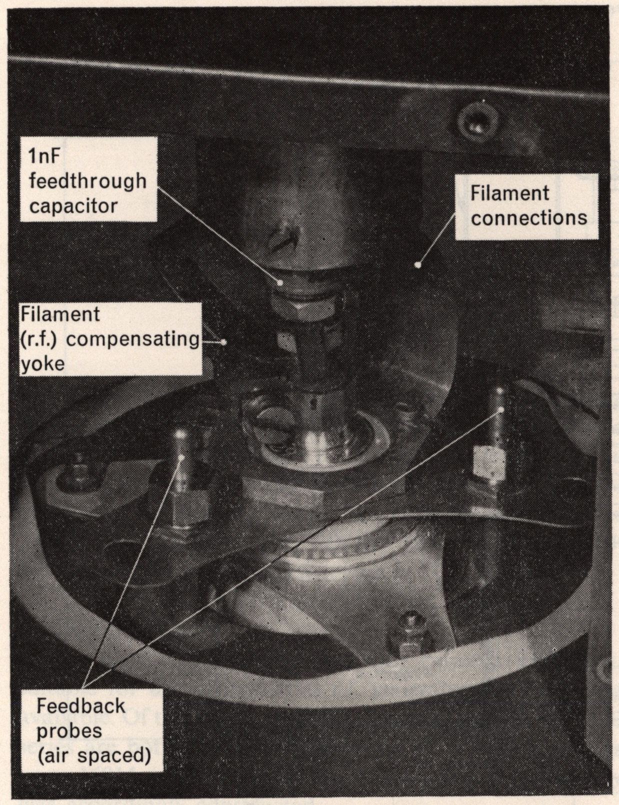
1nF feedthrough capacitor
Filament connections
Filament (r.f.) compensating yoke
Feedback probes (air spaced)

Fig. 4—Details of grid line, filament connectors and feedback system

struction of the grid-filament line is evident, for Fig. 2 shows it positioned coaxially with the main cavity and the valve. The resonant frequency of this line is not critical while it is 100kc/s to 1Mc/s below that of the tank circuit.

In the course of preliminary trials, it was found that the anode efficiency could be greatly influenced by the positioning of the filament decoupling capacitor. If it is not possible to mount this component directly on to and between the filament connectors, the lead lengths between the decoupling capacitor and the filament clamps must be so dimensioned that an optimum r.f. balance of the filament is achieved. The filament balancing loop between filament decoupling capacitor and filament top connector is seen in Fig. 4. The 'live' filament lead is brought out of the centre conductor of the grid-filament line and continued to a small panel at the back (Fig. 2) where a further filament decoupling capacitor is mounted.

Load Circuit

The 50Ω load cable connecting the matched load to the line circuit is terminated in a capacitive probe facing the valve cooler or supporting cylinder. Its distance to the latter is variable, allowing adjustment from no load to about 10% overload. In Fig. 3 the load probe is shown in approximately half load position. It is separated from the valve by a p.t.f.e. sheet to protect the r.f. load from the valve d.c. potential in the case of accidental maladjustment.

PERFORMANCE

Samples of each valve type have been measured in the circuit. The average figures for each type are given below.

The lower load efficiency of the YD1150, if compared with the YD1160, is due to the circuit losses being in both cases approximately the same, thus representing a greater percentage of the YD1150 anode power.

	YD1150		YD1160		YD1161*		
f_0	165		150		150		Mc/s
V_a	5·0		5·0		5·0		kV
R_g	2·0		1·0		1·0		kΩ
	Loaded	*Unloaded*	*Loaded*	*Unloaded*	*Loaded*	*Unloaded*	
I_a	1·0	0·31	2·0	0·40	2·0	0·40	A
I_g	0·28	0·38	0·46	0·65	0·44	0·63	A
p_a	1·25	—	2·5	—	2·5	—	kW
η_a	75	—	75	—	75	—	%
P_L	3·25	—	7·0	—	6·9	—	kW
η_L	65	—	70	—	69	—	%

*Watercooled Type

Read-out of Bistable Transistor Circuits with Gasfilled Diodes ZA1004

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This article gives a detailed analysis of a simple read-out circuit using two gasfilled indicator diodes ZA1004 which are controlled by the small output voltages that can be taken from low-voltage transistor bistable multivibrators.

The general formulae on which the design of the read-out circuit should be based are given for three modes of operation. In addition, in order to facilitate circuit design, a graphical procedure is described, by means of which the required value of the current flowing through the read-out diodes under worst-case conditions can easily be determined. By means of two simple formulae the permissible limits of the auxiliary supply voltage and required series resistor can then readily be evaluated for each mode of operation.

INTRODUCTION

In many transistor logic circuits, particularly bistable multivibrators, it is convenient if the state of the circuits can be read out with an inexpensive device. Filament lamps are unsuitable for this purpose in view of their relatively high current drain and short life. In this respect glow-discharge diodes offer a far better solution. Although by their nature these diodes require a higher operating voltage than is normally available in transistor circuits it is shown in this article that such tubes, fed from a simple auxiliary supply source, can be entirely satisfactorily controlled by the small signals which are obtained from transistors.

The electrical characteristics of the diodes themselves must conform to stringent requirements. The voltage at which the glow discharge is initiated (the ignition voltage, V_{ign}), the voltage produced across the diodes once this discharge is established (the maintaining voltage, V_m) and the voltage at which the discharge ceases (the extinction voltage, V_{ext}) lie close together and remain extremely constant during the life of the tube. In addition, the differences between the characteristics of individual diodes are very small.

The ZA1004 indicator diode has the following outstanding properties:

(1) The molybdenum cathode is kept clean by the continuous sputtering process which takes place over the whole cathode surface during operation. This results in high stability of the characteristics during life: the characteristics remain within the ratings for 15 000 operating hours.

(2) The molybdenum cathode is not affected by the ambient light. The electrical characteristics of the tube thus remain unchanged, whether the tube is used in total darkness or in bright sunlight.

(3) The molybdenum cathode ensures a long tube life and is not affected by temporary overloads.

(4) The diode produces a clearly visible glow discharge for direct observation. The anode is made of gauze so that the glow discharge is visible not only at the end of the tube, but also at the sides.

(5) The size of the meshes of the gauze is such that during operation hardly any sputtered cathode material is deposited on the envelope. As a result, there is no significant decrease of the emitted light during life.

(6) Temperature changes have hardly any effect; the tube operates satisfactorily between temperatures of −55°C and +70°C, the temperature coefficient of the ignition and maintaining voltage being less than 0·02% per deg C.

(7) The tube is of small dimensions (length 25mm max., diameter 6mm approx.), and because of its shape and because it is provided with flying leads, it can easily be mounted in printed circuits.

DESIGN CONSIDERATIONS

Throughout this article, superscripts are used to indicate extreme values of voltages, currents and resistances. Thus $\hat{V}$, $\hat{I}$ and $\hat{R}$ indicate the maximum values of V, I and R, which may be further qualified by subscripts. Similarly $\check{V}$, $\check{I}$ and $\check{R}$ indicate minimum values.

In the current range for which the read-out diode ZA1004 has been designed, it operates in the 'abnormal glow discharge' region. In this region both the lower-limit and the upper-limit maintaining voltage, $\check{V}_m$ and $\hat{V}_m$, rise with the current I_k flowing through the diode. Since the lower-limit ignition voltage $\check{V}_{ign}$ and the lower-limit extinction voltage $\check{V}_{ext}$ drop slightly with temperature, these tube parameters also depend on I_k but to a smaller

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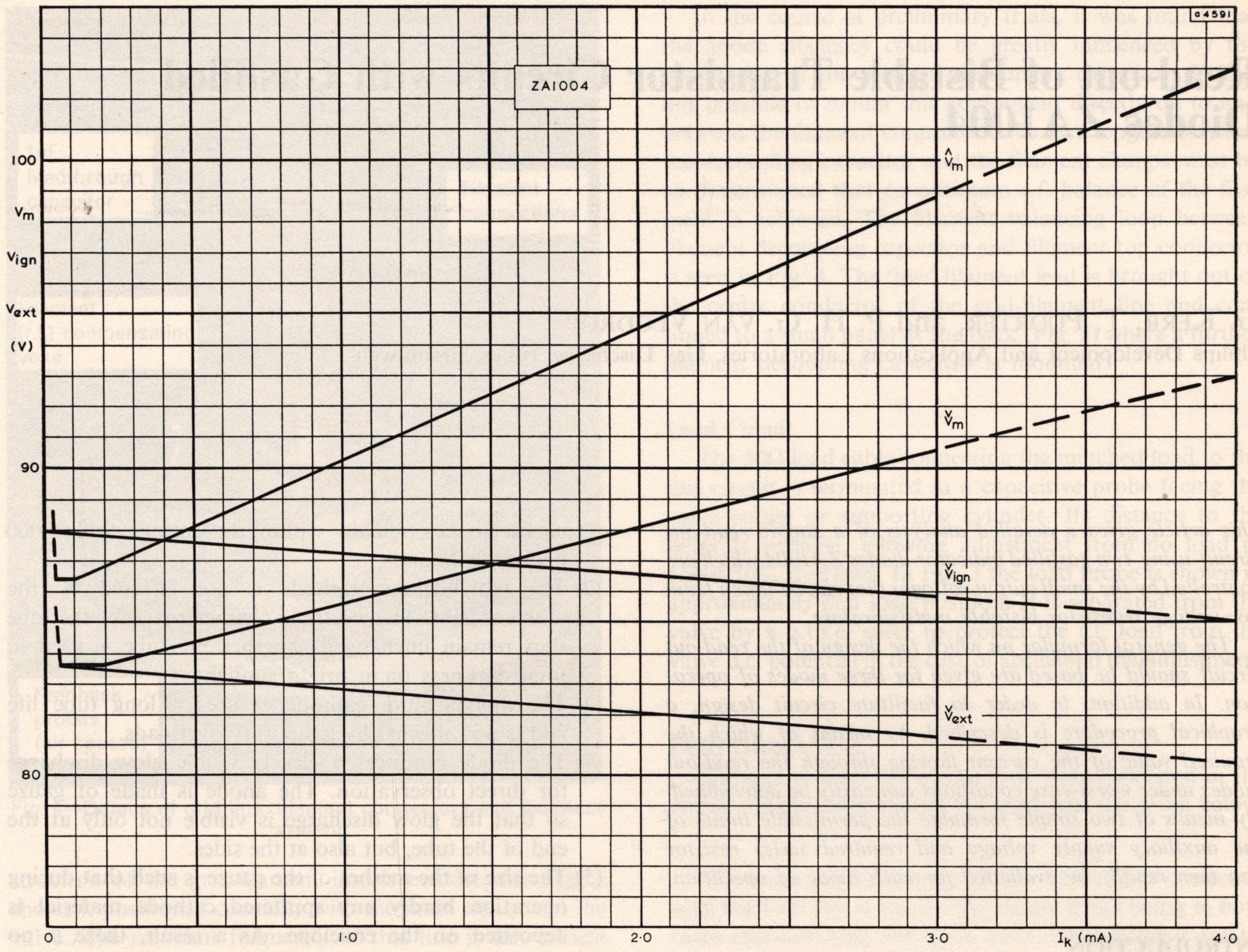


Fig. 1—Values of $\check{V}_m$, $\hat{V}_m$, $\check{V}_{ign}$ and $\check{V}_{ext}$ of gas-filled diode ZA1004 as functions of the cathode current I_k

extent. In Fig. 1 the values of $\check{V}_m$, $\hat{V}_m$, $\check{V}_{ign}$ and $\check{V}_{ext}$ are plotted as functions of the current I_k flowing through the tube.

If the diodes are operated with a pulsating current obtained from a full-wave rectifier (see page 22), the value of $I_{k(m)}$ can be derived from the peak value $I_{k(p)}$ (on which the design must be based) by means of the graph shown in Fig. 2, which gives the ratio $I_{k(m)}/I_{k(p)}$ as a function of the peak value of the supply voltage $V_{b(p)}$.

Since the ZA1004 is essentially an on-off device, its incorporation in the circuit is fundamentally simple. It will nevertheless be clear that the small voltage difference available for switching from the conducting to the non-conducting state and vice versa makes it necessary to choose the component values with care, so as to make allowance for spreads in supply voltage, component values and characteristics.

Fig. 3a shows the basic read-out circuit. The diodes D_1 and D_2 are controlled by the two transistors TR_1 and TR_2, which form part of a bistable multivibrator. The transistors are fed from the common supply source $V_{b'c}$ and the diodes from the auxiliary supply source V_b, via the common resistor R_S.

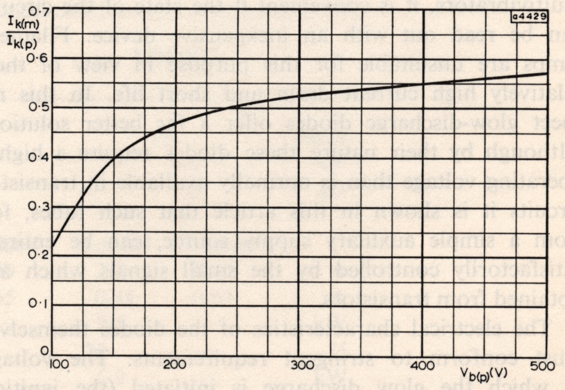


Fig. 2—Ratio $I_{k(m)}/I_{k(p)}$ as a function of the peak value of the supply voltage $V_{b(p)}$

The supply voltage V_b is always taken to be well above the ignition voltage of the read-out diodes, for example 200V. The value of the required operating current I_k is in the first instance determined by the control voltage available from the transistor circuit; it can be given the desired value by suitable choice of V_b and R_S. These must in any case be chosen so that the continuous mean diode current lies within the recommended range of 0·4 to 2·0mA.

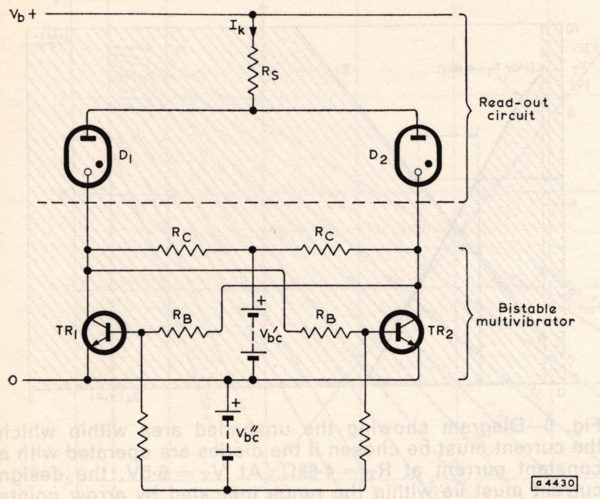


Fig. 3a—Basic circuit of a bistable multivibrator unit with n-p-n transistors to which the read-out circuit is connected

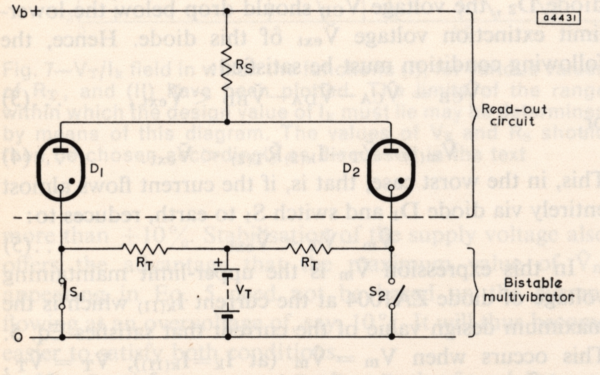


Fig. 3b—Equivalent circuit of Fig. 3a in which the transistors of the multivibrator are represented as switches

Correct operation of the circuit implies that during read-out the diode corresponding to the bottomed transistor is conducting, whilst the other diode is non-conducting. The circuit must therefore be so designed that the following two possibilities are ruled out:

(i) the read-out diode corresponding to the bottomed transistor being non-conducting, while the other diode alone is conducting

(ii) both read-out diodes conducting simultaneously.

It will be shown in the following sections what conditions the diode current I_k must satisfy to ensure that these two possibilities are indeed ruled out, and correct operation can be relied upon. It should be noted in this connection that it is immaterial whether the read-out diodes follow the changes of state of the multivibrator during its operation or not. In practice it is only of interest to read out its final static state.

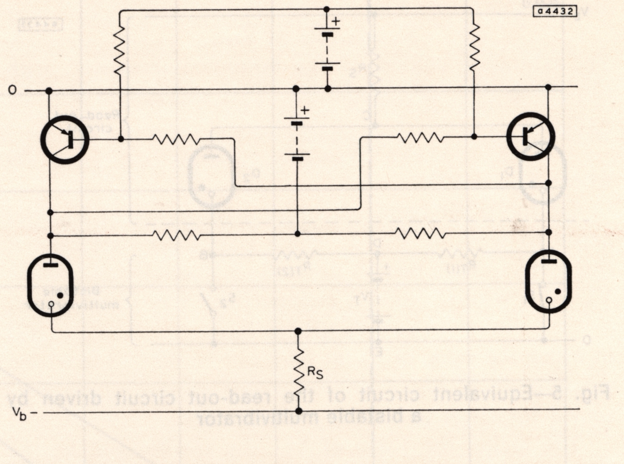


Fig. 4—Circuit similar to that shown in Fig. 3a, but with p-n-p transistors instead of n-p-n transistors

For the sake of clarity the circuit diagram shown in Fig. 3a has been replaced in Fig. 3b by an equivalent circuit in which the transistors TR_1 and TR_2 are represented by ideal switches, S_1 and S_2. In this equivalent circuit, the voltage source $V_{b'c}$ and resistors R_C are replaced by V_T and R_T respectively.

The voltage V_T is smaller than $V_{b'c}$, since it represents the no-load collector voltage measured on the cut-off transistor reduced by the saturation voltage of the bottomed transistor. It should be further recognised that the resistance R_T is smaller than R_C, since R_T should be considered as the output resistance as measured at the collector of the cut-off transistor. The value of R_T may be put approximately equal to $R_C R_B/(R_C+R_B)$. In Fig. 3a the emitters of the transistors, which must be of the n-p-n type, are connected to the common negative supply line. Fig. 4 shows a circuit in which p-n-p transistors are used instead; the emitters are connected to the common positive supply line. The operation of these two circuits is fundamentally the same.

The minimum value of the voltage by which the ZA1004 can be controlled reliably depends on the nature of its auxiliary supply voltage V_b. One of the following three modes of operation can be chosen:

(1) The diodes can be operated with a constant direct current. This requires a minimum control voltage of approximately 5V.

(2) The diodes may be operated with a pulsating direct current obtained from a full-wave rectifier. For this a minimum control voltage of 4·5V is required.

(3) The diodes may be operated with a periodically varying direct current of special waveform. This method requires a minimum control voltage of 3V.

These three modes of operation are discussed separately below.

OPERATION WITH CONSTANT D.C.

It is first necessary to establish what condition must be satisfied to rule out possibility (i), in which only the wrong diode (associated with the cut-off transistor) is conducting.

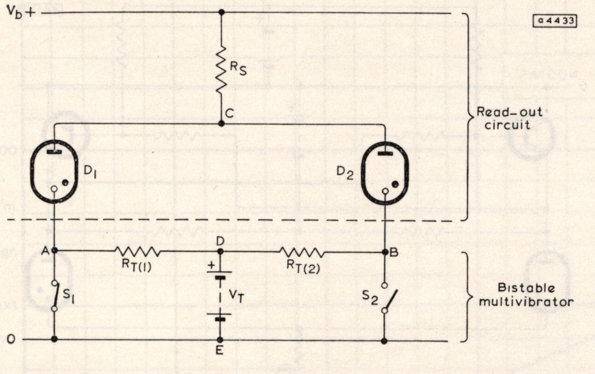


Fig. 5—Equivalent circuit of the read-out circuit driven by a bistable multivibrator

Assume, for example, switch S_1 (Fig. 5) to be closed (corresponding to transistor TR_1 in Fig 3a being bottomed), but the wrong read-out diode D_2 to be conducting. Should this occur, it is clear that a current originating from V_b is flowing via R_S, D_2, $R_{T(2)}$ and voltage source V_T to earth and, in addition, a current originating from V_T is flowing via $R_{T(1)}$ and S_1 to earth. The voltage produced across diode D_1 is thus:

$$V_{CA} = V_{CB}+V_{BD}+V_{DA} = V_m+I_kR_{T(2)}+V_T. \qquad \ldots(1)$$

To ensure that the correct diode is ignited even under extreme conditions, the following general inequality must be satisfied:

$$\check{V}_m+I_{k(I)}\check{R}_T+\check{V}_T' > \hat{V}_{ign}. \qquad \ldots(2)$$

In this expression $I_{k(I)}$ denotes the minimum value of the current on which the design must be based to ensure that Eq. 2 is satisfied under the most unfavourable condition, that is if V_m, R_T and V_S have their lower-limit values $\check{V}_m$ (at $I_k = I_{k(I)}$), $\check{R}_T$ and $\check{V}_S$ respectively, whilst R_S has its upper-limit value $\hat{R}_S$. The voltage $\check{V}_T'$ denotes the lowest control voltage which might be present at $R_T = \hat{R}_T$, in accordance with the second term of the left-hand side of Eq. 2. This distinguishes it from the lower-limit control voltage V_T, the calculation of which should be based on the assumption that the transistor leakage current flows through R_T having its upper-limit value $\hat{R}_T$.

Since for a given multivibrator circuit both R_T and V_T are fixed and $\hat{V}_{ign}$ is specified in the tube data, all that is required is to ensure that the current supplied by the auxiliary voltage source is high enough to ensure Eq. 2 is satisfied in the event of only the wrong diode being conducting. This point is discussed further on page 21.

Since in Eq. 2, not only the term $I_{k(I)}R_T$, but also V_m increases with current, the higher this current, the easier it will be to satisfy Eq. 2. This is illustrated by the following example.

Assume $\check{V}_T'$ to be 6·5V and $\check{R}_T$ to be 4·0kΩ. Under these conditions, at $I_{k(I)} = 0{\cdot}5$mA, the left-hand side of Eq. 2 would become:

$$84{\cdot}25+0{\cdot}5\times4{\cdot}0+6{\cdot}5 = 92{\cdot}75\text{V},$$

in which case the wrong diode will not ignite the correct one with certainty, since the upper-limit ignition voltage $\hat{V}_{ign} = 93$V. The current $I_{k(I)}$ should therefore be raised to, say, 0·55mA, which gives 93·1V for the left-hand side

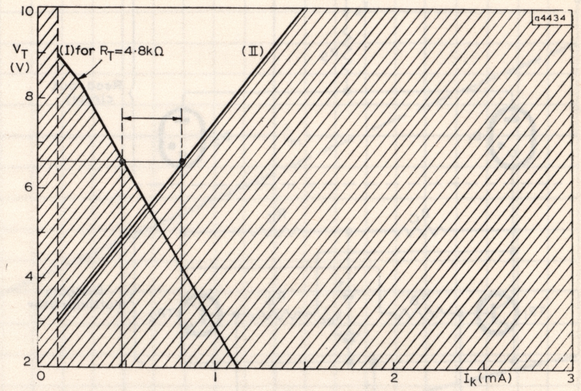


Fig. 6—Diagram showing the unshaded area within which the current must be chosen if the diodes are operated with a constant current at $R_T = 4{\cdot}8$kΩ. At $V_T = 6{\cdot}5$V, the design current must lie within the range indicated by arrow points

of Eq. 2. Ignition of the correct diode is then ensured as soon as the statistical ignition delay, the average value of which is 50ms, has elapsed.

Now we can derive the condition at which possibility (ii), when both diodes conduct simultaneously, is also ruled out. For this purpose, to ensure extinction of diode D_2, the voltage V_{CB} should drop below the lower-limit extinction voltage $\check{V}_{ext}$ of this diode. Hence, the following condition must be satisfied:

$$\hat{V}_{CB} = \hat{V}_{CA}-\check{V}_{DA}-\check{V}_{BD} < \check{V}_{ext}, \qquad \ldots(3)$$

or,

$$\hat{V}_{m(1)}-\check{V}_T-I_{k(2)}\check{R}_{T(2)} < \check{V}_{ext}. \qquad \ldots(4)$$

This, in the worst case, that is, if the current flows almost entirely via diode D_1 and switch S_1 to earth, reduces to:

$$\hat{V}_m-\check{V}_T < \check{V}_{ext}. \qquad \ldots(5)$$

In this expression $\hat{V}_m$ is the upper-limit maintaining voltage of diode ZA1004 at the current $I_{k(II)}$ which is the maximum design value of the current that satisfies Eq. 5. This occurs when $V_m = \hat{V}_m$ (at $I_k=I_{k(II)}$), $V_T = \check{V}_T$, $V_b = \hat{V}_b$ and $R_S = \check{R}_S$. $\check{V}_{ext}$ is the lower-limit extinction voltage, also at this value of $I_{k(II)}$.

Since $\hat{V}_m$ increases and $\check{V}_{ext}$ decreases with I_k, this current has an upper limit below which condition (ii) is also ruled out. If, in the example discussed above, $\check{V}_T = \check{V}_T'-0{\cdot}1 = 6{\cdot}4$V, the condition is thus imposed that:

$$\hat{V}_m-\check{V}_{ext} < \check{V}_T = 6{\cdot}4\text{V}.$$

Examination of the graph shown in Fig. 1 reveals that the current should not exceed 0·78mA to comply with this condition; $\hat{V}_m$ and $\check{V}_{ext}$ are then 89·3V and 82·9V respectively.

In the example discussed, correct operation of the circuit is thus ensured, provided that $I_{k(I)} > 0{\cdot}55$mA and $I_{k(II)} < 0{\cdot}78$mA, according to Eqs. 2 and 5.

If the margin between $I_{k(I)}$ and $I_{k(II)}$ is very small, it may be necessary to stabilise the supply voltage to the read-out diodes. In that case, the value of $I_{k(I)}$ (and hence also of $\check{V}_m$) appearing in Eq. 2 need not be based on an undervoltage of, say, 10%. It should be recognised in this respect, that almost the entire supply voltage variation appears across resistor R_S, so that a supply voltage variation of, say, ±10% results in a variation of I_k of

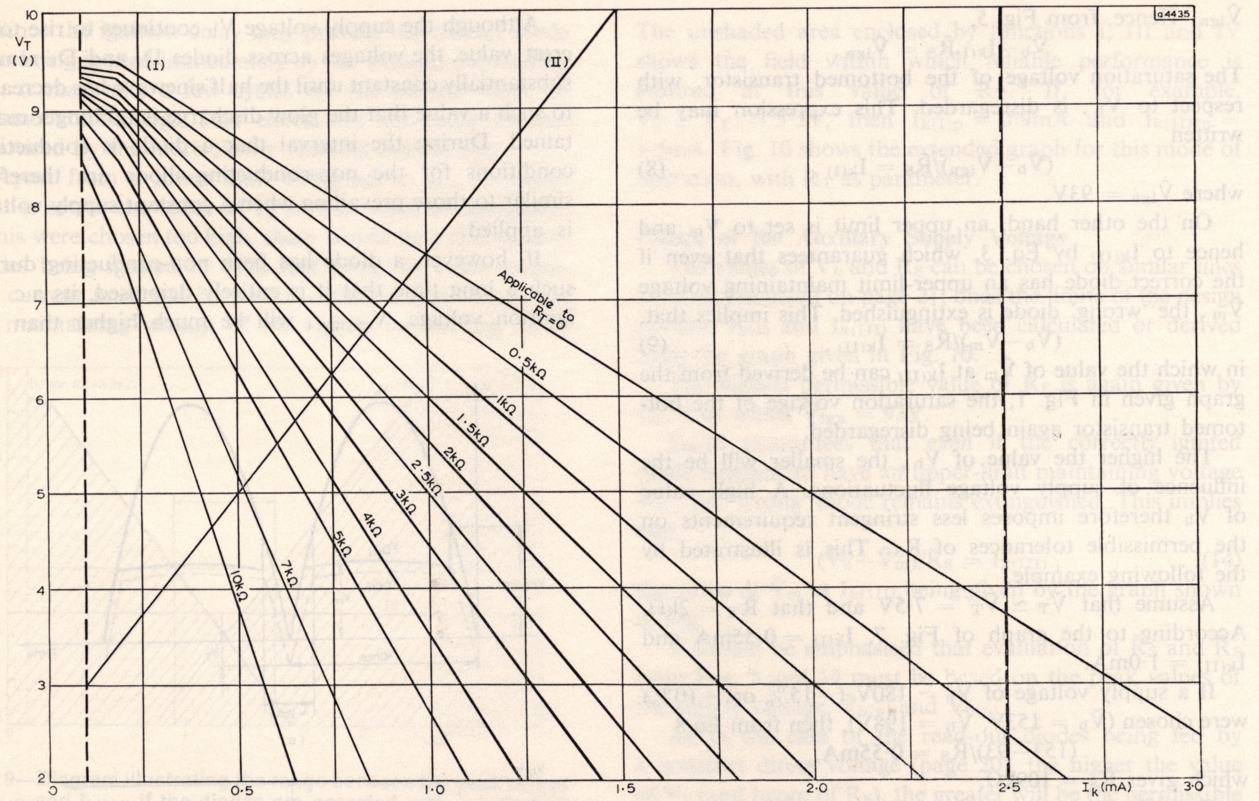


Fig. 7—V_T/I_k field in which the functions (I), for various values of R_T, and (II) have been plotted. The limits of the range within which the design value of I_k must lie may be determined by means of this diagram. The values of V_S and R_S should then be chosen accordingly as discussed in the text

more than ±10%. Stabilisation of the supply voltage also offers the advantage that the maximum value of $\hat{V}_m$ appearing in Eq. 5 need not be based on the current flowing at an overvoltage of, say, 10%. It will thus become easier to satisfy both conditions.

If no satisfactory compromise can be found between Eqs. 2 and 5, a solution might be to design the multivibrator so that R_T has a value in excess of 4·0kΩ (see Eq. 1). However, if germanium transistors are employed, their collector resistors should not have too large a value, because of their increased collector leakage current at high ambient temperatures. Alternatively, a different mode of operation (see below, pages 22 and 23) may be chosen.

Graphical Design Procedure

The design of the read-out circuit is greatly facilitated by using the graphical method explained below. From a single diagram the necessary information can be derived to enable a read-out circuit to be designed for a given multivibrator circuit, while still giving reliable performance under worst-case conditions.

In consulting the diagrams the starting point is always the available control voltage V_T supplied by the multivibrator and the value of the resistance R_T.

Eq. 2, which must be satisfied if the read-out circuit is supplied by a constant direct voltage, can be represented graphically by plotting in the V_T-I_k field (Fig. 6), for a given value of $\check{R}_T$, the function

$$V_T = \hat{V}_{ign}-\check{V}_m-I_{k(I)}\check{R}_T. \qquad \ldots(6)$$

In this equation $\check{V}_m = f(I_{k(I)})$ is derived from the corresponding curve in Fig. 1, and $\hat{V}_{ign} = 93$V. Curve (I), thus obtained, represents the minimum design current, $I_{k(I)}$, as a function of V_T, valid at $R_T = 4{\cdot}8$kΩ. Similarly, Eq. 5 can be represented in this field by plotting the function

$$V_T = \hat{V}_m-\check{V}_{ext}. \qquad \ldots(7)$$

Here, $\hat{V}_m$ and $\check{V}_{ext}$ are both derived from the corresponding curves given in Fig. 1. In this way function (II) representing the maximum design current, $I_{k(II)}$, as a function of V_T, is obtained.

The unshaded area enclosed by curves (I) and (II) thus represents the field within which reliable operation is ensured at $R_T = 4{\cdot}8$kΩ.

The use of this graph will be illustrated by assuming $V_T \simeq V_T'$ to be 6·5V. The graph shows that $I_{k(I)} = 0{\cdot}48$mA and $I_{k(II)} = 0{\cdot}75$mA. A suitable value for I_k may thus be 0·6mA, leaving a margin in both directions, which is necessary to make allowance for variations of the supply voltage V_S and spread in the resistor values.

The graph of Fig. 6 can be extended by plotting, instead of curve (I), applicable to $R_T = 4{\cdot}8$kΩ, a family of curves for various values of R_T. The diagram of Fig. 7 is thus obtained.

Choice of the Auxiliary Supply Voltage

Once the minimum and maximum design values of the current, $I_{k(I)}$ and $I_{k(II)}$, have been calculated or ascertained by means of the graph given in Fig. 7, we still need to choose V_b and R_S accordingly.

V_b and R_S must be so chosen that the voltage produced across the diode exceeds the upper-limit ignition voltage

$\hat{V}_{ign}$. Hence, from Fig. 5,

$$\check{V}_b - I_{k(I)}\hat{R}_S = \hat{V}_{ign}\,.$$

The saturation voltage of the bottomed transistor, with respect to V_b, is disregarded. This expression may be written

$$(\check{V}_b - \hat{V}_{ign})/\hat{R}_S = I_{k(I)}\,, \qquad \ldots(8)$$

where $\hat{V}_{ign} = 93$V.

On the other hand, an upper limit is set to V_m and hence to $I_{k(II)}$ by Eq. 3, which guarantees that even if the correct diode has an upper-limit maintaining voltage $\hat{V}_m$, the 'wrong' diode is extinguished. This implies that,

$$(\hat{V}_b - \hat{V}_m)/\check{R}_S = I_{k(II)}\,, \qquad \ldots(9)$$

in which the value of $\hat{V}_m$ at $I_{k(II)}$ can be derived from the graph given in Fig. 1, the saturation voltage of the bottomed transistor again being disregarded.

The higher the value of V_b, the smaller will be the influence of supply voltage fluctuations. A high value of V_b therefore imposes less stringent requirements on the permissible tolerances of R_S. This is illustrated by the following example.

Assume that $\check{V}_T \simeq \check{V}_T' = 7{\cdot}5$V and that $\check{R}_T = 2$kΩ. According to the graph of Fig. 7, $I_{k(I)} = 0{\cdot}55$mA and $I_{k(II)} = 1{\cdot}0$mA.

If a supply voltage of $V_b = 180$V (-15% or $+10\%$) were chosen ($\check{V}_b = 153$V, $\hat{V}_b = 198$V), then from Eq.8

$$(153-93)/\hat{R}_S = 0{\cdot}55\text{mA},$$

which gives $\hat{R}_S = 109$kΩ.

From Eq.9

$$(198-90)/\check{R}_S = 1{\cdot}0\text{mA}.$$

Thus $\check{R}_S = 108$kΩ.

The value of resistor R_S would thus have to lie within the limits of 108kΩ and 109kΩ, which is obviously impracticable.

If in the above example the supply voltage were raised to $V_S = 250$V(-15% or $+10\%$), calculation shows that a resistor R_S of 200kΩ with an end-of-life tolerance of ±7% could be used.

OPERATION WITH FULL-WAVE UNSMOOTHED RECTIFIED SUPPLY

In the mode of operation discussed below, the read-out circuit is fed from a voltage having the waveform as shown in Fig. 8a. This mode of operation has the advantage that whichever tube happens to be conducting, it will be extinguished as soon as the supply voltage drops below the extinction voltage. This occurs automatically twice in each cycle of the mains supply. It is therefore unnecessary to consider, in this case, the possibility of a wrongly conducting diode not being extinguished. In other words, there is no need to take into account Eq. 5.

The operation of the circuit can be followed by reference to Fig. 8b. Assume switch S_1 to be closed and S_2 to be open, diode D_1 to be conducting and diode D_2 to be non-conducting. The cathode of diode D_2 will then be positively biased by a voltage equal to V_T.

In normal operation, read-out diode D_1 will then be reignited periodically each time the sinusoidal supply voltage has reached the 'a.c. re-ignition voltage, $V_{re\text{-}ign}$' corresponding to this mode of operation. The voltage across this diode then suddenly drops to the slightly lower maintaining value.

Although the supply voltage V_b continues to rise to its crest value, the voltages across diodes D_1 and D_2 remain substantially constant until the half sinewave has decreased to such a value that the glow discharge is no longer maintained. During the interval that a diode is conducting, conditions for the non-conducting diode are therefore similar to those prevailing when a constant supply voltage is applied.

If, however, a diode has been non-conducting during such a long time that it is entirely deionised, its a.c. re-ignition voltage, $V_{re\text{-}ign}$, will be much higher than the published upper-limit value of 101V. If the multivibrator then changes its state, the diode which was previously conducting is certainly ignited first during the next half cycle of the supply voltage although it is positively biased because its transistor is cut off.

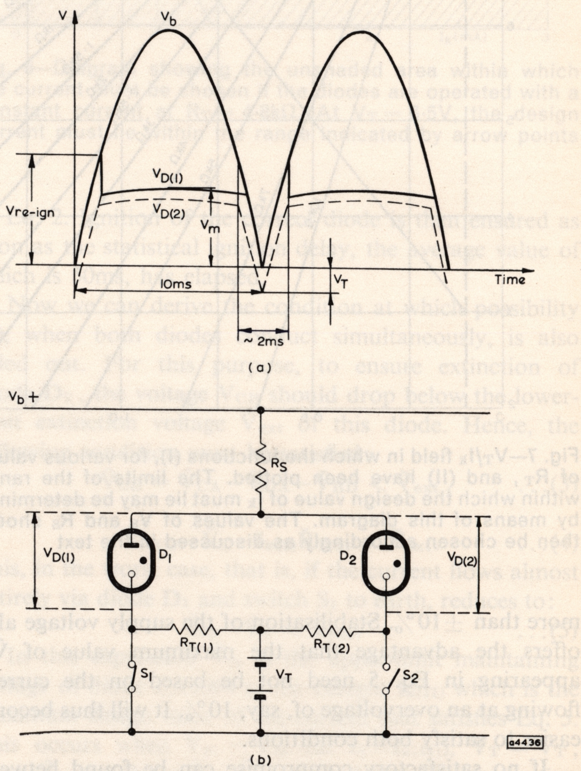


Fig. 8(a)—Various voltages occurring as functions of time in read-out circuit of Fig. 8(b) operated with a pulsating current obtained from a full-wave rectifier

It is stressed in this connection that the supply voltage must be obtained from a 50c/s to 60c/s single-phase full-wave rectifier. Only in that case will the interval between the consecutive half cycles be short enough to ensure ignition of a diode at the published a.c. re-ignition voltage $\hat{V}_{re\text{-}ign}$ of 101V.

To allow the correct diode to become conducting under these conditions it is essential that, during the interval in which the wrongly ignited diode is still conducting, the peak voltage then appearing across the correct diode is so high that the upper-limit d.c. ignition voltage $\hat{V}_{ign}$ is exceeded. In other words, in this case also, a lower limit is set to the peak value of I_k by Eq. 2. The peak value of $\hat{V}_m$ can again be derived from the graph shown in Fig. 1.

Since the ignition delay may prevent the correct diode from being ignited immediately by the wrongly conducting one, it may take a few cycles for the correct diode to become conducting. This effect will obviously remain unnoticed by the more slowly reacting human eye.

Apart from a lower limit being set to I_k, an upper limit is set to the value of I_k in this mode of operation. If this were chosen too high, there would be a risk of both diodes being ignited in each half cycle. In other words, this current must not be raised to such a high value that the maintaining voltage of the rightly conducting diode can cause the wrong diode to ignite, even though the latter is biased. This condition, to be satisfied by the peak value of the steady-state current, can be expressed by,

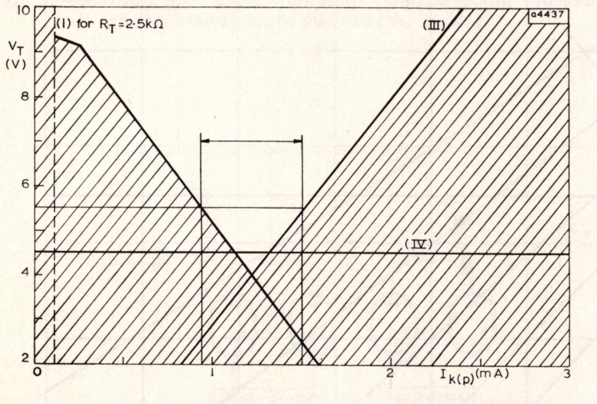


Fig. 9—Diagram illustrating the range between the peak values of $I_{k(I)}$ and $I_{k(III)}$ if the diodes are operated with a pulsating current taken from a full-wave rectifier at $V_T = 5{\cdot}5$V and $R_T = 2{\cdot}5$kΩ

$$\hat{V}_m - \check{V}_T < \check{V}_{ign}\,, \qquad \ldots(10)$$

in which $\hat{V}_m$ and $\check{V}_{ign}$ must now be based on $I_k = I_{k(III)}$, that is, the maximum design current at which Eq. 10 is still satisfied. This occurs when $V_m = \hat{V}_m$ (at $I_k = I_{k(III)}$), $V_T = \check{V}_T$, $V_b = \hat{V}_{b(p)}$ and $R_S = \check{R}_S$.

To ensure that the correct diode, once it has been ignited, will reignite with certainty during the next few half cycles of the auxiliary supply voltage, the bias voltage V_T must obviously exceed the difference between the upper-limit a.c. re-ignition voltage $\hat{V}_{re\text{-}ign} = 101$V and the lower-limit a.c. re-ignition voltage $\check{V}_{re\text{-}ign} = 96{\cdot}5$V. The following additional condition is thus imposed on $\check{V}_T$ in this mode of operation,

$$\check{V}_T > \hat{V}_{re\text{-}ign} - \check{V}_{re\text{-}ign} = 4{\cdot}5\text{V}. \qquad \ldots(11)$$

Graphical Design Procedure

For this mode of operation, it is also possible to ascertain the permissible values of I_k which ensure reliable performance under worst-case conditions from a single diagram, similar to that given in Fig. 7.

For this purpose it is necessary to plot in the $V_T - I_k$ field, in addition to function (I) (see page 21), two further functions. These, derived from Eqs. 10 and 11, are

$$V_T = \hat{V}_m - \check{V}_{ign}\,, \qquad \ldots(12)$$

(in which $\hat{V}_m$ and $\check{V}_{ign}$ are derived from the corresponding curves in Fig. 1), and,

$$V_T = 4{\cdot}5\text{V}. \qquad \ldots(13)$$

These are called functions III and IV.

Fig. 9 shows the graph thus obtained for $R_T = 2{\cdot}5$kΩ. The unshaded area enclosed by functions I, III and IV shows the field within which reliable performance is ensured at this value of R_T. If, for example, $V_T \simeq V_T' = 5{\cdot}5$V, then $I_{k(I)p} = 0{\cdot}9$mA and $I_{k(III)p} = 1{\cdot}5$mA. Fig. 10 shows the extended graph for this mode of operation, with R_T as parameter.

Choice of the Auxiliary Supply Voltage

The values of V_b and R_S can be chosen on similar lines to those discussed on page 21, once the limits of the design current, $I_{k(I)}$ and $I_{k(III)}$ have been calculated or derived from the graph given in Fig. 10.

The highest permissible value of R_S is again given by Eq. 8 in which $\hat{V}_{ign} = 93$V.

Eq.10 guarantees that even if the correctly ignited diode happens to have an upper-limit maintaining voltage $\hat{V}_m$, the 'wrong' diode remains extinguished. This implies that,

$$(\hat{V}_b - \hat{V}_m)/\check{R}_S = I_{k(III)}\,, \qquad \ldots(14)$$

the value of $\hat{V}_m$ at $I_{k(III)}$ being given by the graph shown in Fig. 1.

It should be emphasised that evaluation of $\hat{R}_S$ and $\check{R}_S$ from Eqs. 8 and 14 must be based on the peak values of the supply voltage, $\check{V}_{b(p)}$ and $\hat{V}_{b(p)}$.

As in the case of the read-out diodes being fed by a constant direct voltage (page 20), the higher the value of V_b (and hence of R_S), the greater will be the permissible tolerances of R_S and acceptable fluctuations of V_b. This is of particular interest here, because with this mode of operation it is difficult to stabilise the pulsating supply voltage.

OPERATION WITH PERIODICALLY VARYING D.C. OF SPECIAL WAVEFORM

By choosing one of the methods of operation discussed in this section, in which the imposed conditions are satisfied consecutively, it is possible to drive the read-out circuit with a control voltage of only 3V. For this purpose any one of four procedures outlined below can be followed.

(a) The read-out circuit is operated at a (low) steady-state current which does not exceed the value given by the previously derived Eq. 5. On this steady-state current, a positive-going pulse is superimposed, the peak value of which exceeds that given by Eq. 2, thus ensuring ignition of the correct diode. Subsequently, the current drops back to the steady-state value, thus ensuring extinction of a wrongly conducting diode.

(b) The read-out circuit is operated at a steady-state current which fulfils Eq. 2, so that the correct diode is ignited; on this steady-state current a negative-going pulse of such amplitude is superimposed that Eq. 5 is satisfied, thus ensuring extinction of a wrongly conducting diode. However, apart from Eq. 2, Eq. 10 also determines the steady-state current; this current must not be raised to so high a value that the maintaining voltage of the rightly conducting diode can cause the wrong diode to ignite, even though the latter is biased.

(c) The read-out circuit is operated at a value of the steady-state current which does not conflict with the maximum value set by Eq. 10. On this steady-state current is

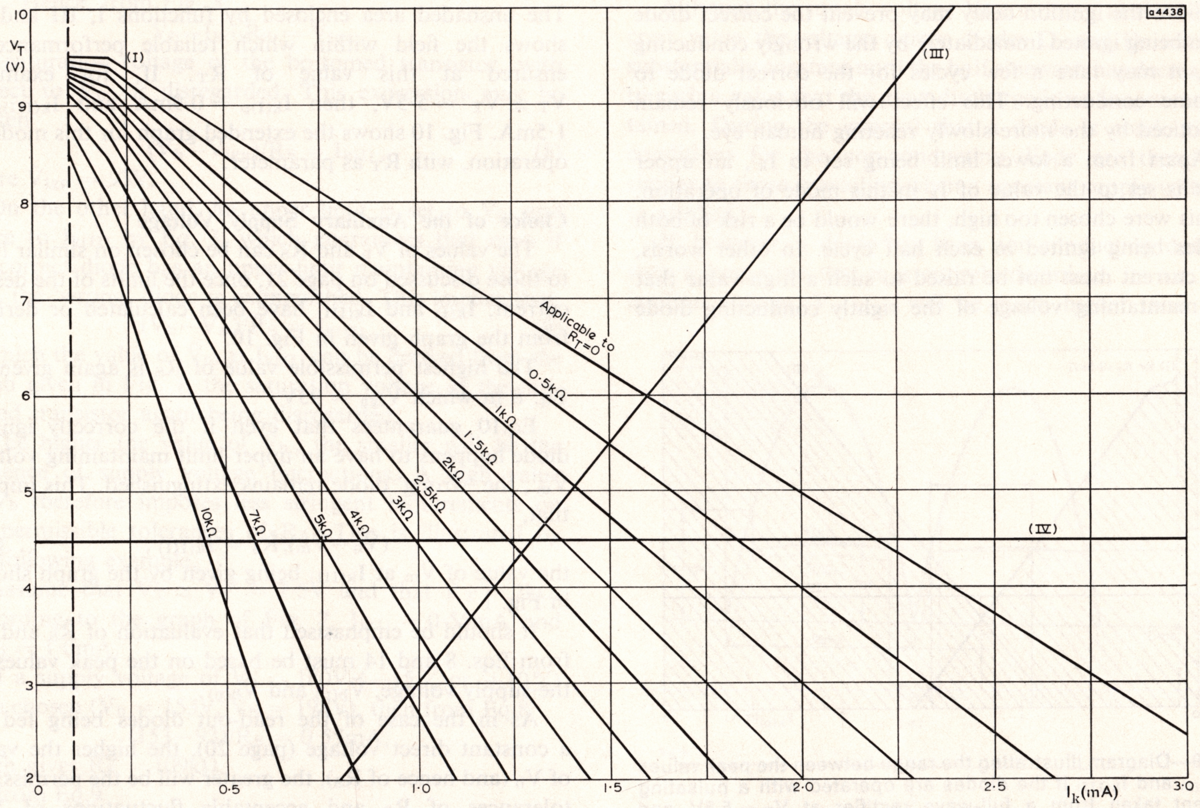


Fig. 10—V_T/I_k field in which the functions (I), for various values of R_T, (III) and (IV) have been plotted. After having derived from this graph the limits of the range within which the design value of I_k must lie, the values of V_S and R_S can be ascertained by means of Eqs. 8 and 14

superimposed a positive-going pulse followed by a negative-going pulse, the amplitudes of these pulses being such that Eqs. 2 and 5 are satisfied in succession. During the positive-going pulse the correct diode is ignited, whilst the subsequent negative-going pulse again ensures extinction of a wrongly conducting diode.

(d) For those cases where it is acceptable to operate the read-out switch manually whenever necessary, the following method is suitable. The steady-state current may assume any value, provided Eq. 2 is satisfied. By depressing the switch, the current is made to drop momentarily to so low a value that Eq. 5 is satisfied, after which it may assume a higher value, provided this does not conflict with Eq. 10. After the multivibrators have been read out, the switch is released.

Figs. 11a, b, c and d show examples of current waveforms for these four methods.

The advantage of method (a) is the low power dissipation. However, since the steady-state current should preferably exceed 0·4mA to ensure a high constancy of the tube characteristics, the advantages of this method are somewhat restricted.

Since with method (b) the steady-state value of the current need not satisfy Eq. 5, it may be given a higher value, but lower and upper limits are set to this current by Eqs. 2 and 10.

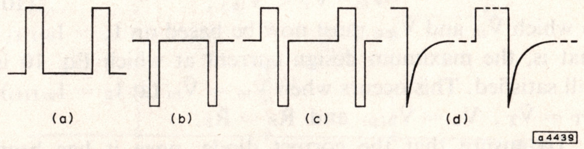


Fig. 11—Schematic representation of the variations of I_k with time if the diodes are operated with a periodically varying direct current of special waveform. (a) Constant steady-state current with superimposed positive-going pulses, (b) constant steady-state current with superimposed negative-going pulses, (c) constant steady-state current with superimposed positive-going pulses followed by negative-going pulses, and (d) suggested variation of the steady-state current if the read-out switch is to be operated manually.

Method (c) differs from method (b) in that only an upper limit (Eq. 10) affects the steady-state current, so that this method offers a wider choice of current value.

The advantages of method (d) are similar to those of method (c), but the necessity of operating the read-out switch manually may not be acceptable.

Circuits for Obtaining the Required Supply Voltage Variations

For method (a) the circuit shown in Fig. 12 may be used. Switch S_1 is normally open. When it is closed, a supply voltage V_b' in excess of V_b is temporarily applied to the read-out diodes, so that I_k is raised to such a value that the correct diodes are certain to be ignited. After the

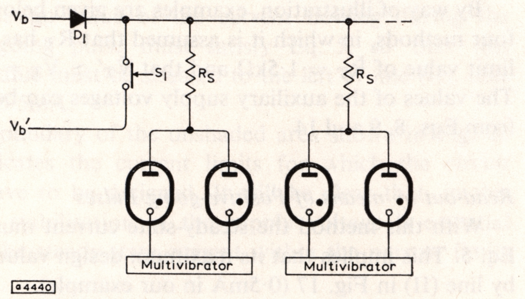


Fig. 12—Method for superimposing positive-going pulses on the steady-state current ($V_b' > V_b$)

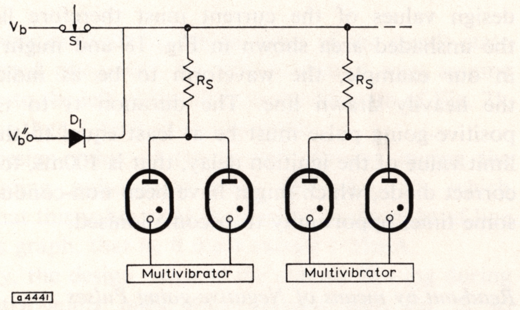


Fig. 13—Method for superimposing negative-going pulses on the steady-state current ($V_b'' < V_b$)

switch is opened again, the current drops back to the steady-state value, at which the circuit can then be read out. The common semiconductor diode D_1 is necessary to permit the line voltage being raised from V_b to V_b'.

Fig. 13 shows a similar circuit for read-out purposes if method (b) is used. Switch S_1, which is normally closed, is temporarily opened to reduce the supply voltage from its normal value V_b to V_b'', at which value wrongly conducting diodes are certain to be extinguished. After the switch is closed again, the circuit can be read out at the steady-state current. In this circuit the common semiconductor diode D_1 is included in the V_b'' line.

Fig. 14 shows a circuit for method (c) by means of which the current can be raised and lowered successively before returning to the steady-state value. Each read-out circuit is here provided with an auxiliary resistor R_S' which must have a high value compared with R_s. During normal operation switch S_1 is in position 1 (the normal resistance in the supply line of each read-out circuit is then R_S in parallel with R_S'). When the switch is temporarily moved to position 2, the supply voltage is raised from its normal value V_b to V_b', thus producing the positive-going current pulse required to ensure that the correct diode is ignited. Advantage is taken of the fact that, during the subsequent return of the switch from position 2 to position 1, the current is forced to flow from V_b through resistor R_S' of high value, thus resulting in the negative-going current pulse required for extinguishing any wrongly ignited diodes. Once the switch has returned to position 1, the

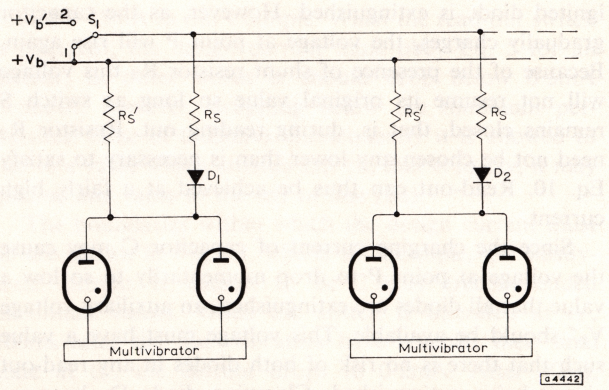


Fig. 14—Method for superimposing a positive-going pulse followed by a negative-going pulse on the steady-state current ($V_b' > V_b$)

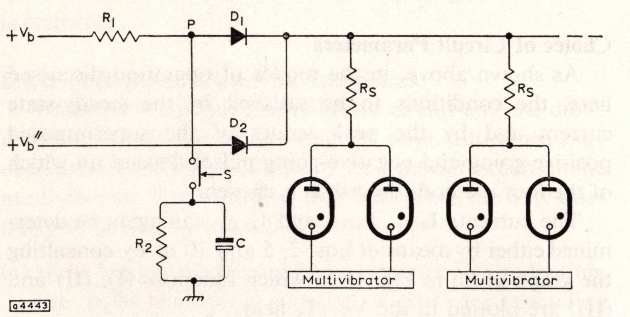


Fig. 15—Recommended circuit for reading-out the multivibrators by means of a manually operated switch ($V_b'' < V_b$)

circuit can be read out at the steady-state current. It is of no importance at all that the current is also temporarily lowered when the switch changes from position 1 to position 2, prior to the positive-going read-out pulse being produced. The semiconductor diodes in series with each read-out circuit have been provided to eliminate the risk of interference between the individual circuits.

With methods (a), (b) and (c) preference will usually be given to an automatically operated relay or to an electronic switching device to achieve the required current variation. It is possible to arrange the supply circuit so that the steady-state current flows continuously and that the variation required for read-out purposes is brought about only when the circuit is to be read out. Alternatively, the supply current can be made to vary at regular intervals. With methods (a) and (c) the repetition frequency should then be chosen such that the short lighting-up of the wrong diode during the occurrence of the positive-going pulse is not annoying.

The minimum time during which the steady-state current should be raised or lowered to ensure proper operation of the read-out circuits is discussed below.

Fig. 15 shows a circuit for operating the read-out switch manually, as in method (d). The read-out diodes are operated at a high steady-state current, say 2mA, supplied by voltage source V_b via the common resistor R_1. As soon as switch S is closed, the charging current of capacitor C will cause the voltage at point P to drop suddenly to a very low value. This ensures that any wrongly ignited diode is extinguished. However, as the capacitor gradually charges, the voltage at point P will rise again. Because of the presence of shunt resistor R_2 this voltage will not resume its original value so long as switch S remains closed, that is, during reading out. Resistor R_2 need not be chosen any lower than is necessary to satisfy Eq. 10. Read-out can thus be achieved at a fairly high current.

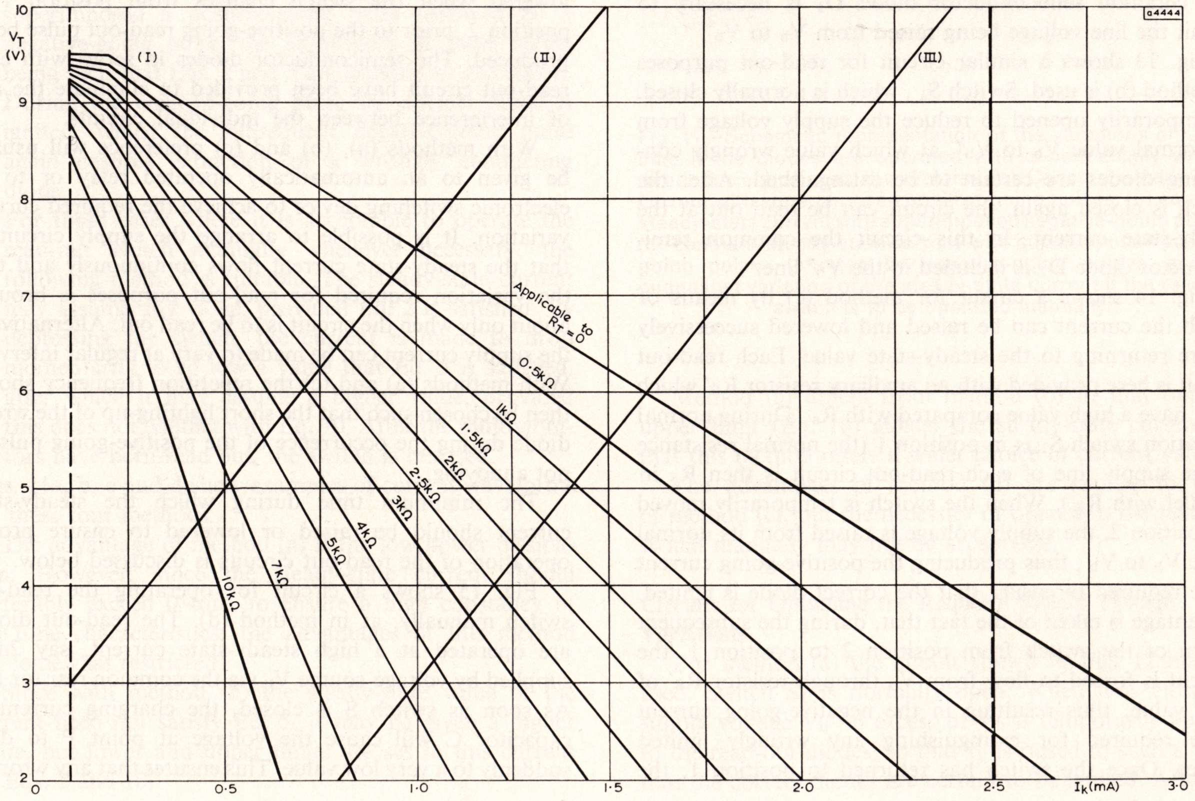
Since the charging current of capacitor C may cause the voltage at point P to drop momentarily to so low a value that all diodes are extinguished, an auxiliary voltage V_b'' should be available. This voltage must have a value such that there is no risk of both diodes in any read-out circuit being extinguished. Clamping diode D_2 becomes conducting only during the time that the potential at point P is lower than V_b''. In that case diode D_1 prevents the current supplied by V_b'' from flowing towards point P.

Choice of Circuit Parameters

As shown above, in the modes of operation discussed here, the conditions to be satisfied by the steady-state current and by the peak values of the superimposed positive-going and negative-going pulses depend on which of the four methods described is chosen.

The currents $I_{k(I)}$, $I_{k(II)}$ and $I_{k(III)}$ can again be determined either by means of Eqs. 2, 5 and 10 or by consulting the graph given in Fig. 16 in which functions (I), (II) and (III) are plotted in the V_T-I_k field.

Fig. 16—Graph showing V_T/I_k field in which the functions (I), for various values of R_T, (II) and (III) have been plotted, thus permitting the designer to ascertain the design values of the steady-state current and superimposed pulses if the read-out circuit is operated with a periodically varying current of special waveform



By way of illustration, examples are given below for the four methods, in which it is assumed that R_T has a lower-limit value of $\check{R}_T = 1{\cdot}5\text{k}\Omega$ and that $\check{V}_T' \simeq \check{V}_T = 5\text{V}$. The values of the auxiliary supply voltages can be derived from Eqs. 8, 9 and 14.

Read-out by means of Positive-going Pulses

With this method the steady-state current must satisfy Eq. 5. This implies that its maximum design value is given by line (II) in Fig. 17 (0·5mA in our example).

The positive-going pulse must satisfy Eq.2. In other words, its minimum design value is given by line (I), and should thus exceed 1·25mA in this example.

The data indicate that the steady-state current should preferably not be chosen below 0·4mA, whilst the peak of the pulse is allowed to reach a value of 3mA. The design values of the current must therefore lie within the unshaded area shown in Fig. 18 and might assume, in our example, the waveform to be as indicated by the heavily drawn line. The duration t_1 to t_2 of the positive-going pulse must be at least equal to the upper-limit value of the ignition delay, that is 100ms, to give the correct diode, which might have been non-conducting for some time, opportunity to become ionised.

Read-out by means of Negative-going Pulses

With this method a lower limit is imposed on the design value of the steady-state current by Eq. 2 and an upper limit by Eq. 10.

In our example the design value of the steady-state current would therefore have to lie within the unshaded areas between lines (I) and (III) in Fig. 19 (page 28), hence between 1·25mA and 1·4mA.

The design value of the current flowing during the negative-going pulse must satisfy Eq. 5, so that this current value must be situated to the left of line (II), that is below 0·5mA.

The boundary of the unshaded area shown in Fig. 20 thus indicates the current limits for which the circuit would have to be designed. It will be clear that, in this example, stabilisation of the supply voltage is essential to keep the steady-state current within the narrow limits indicated.

After the multivibrator has reached its final state, at t_1, an interval t_1 to t_2 of at least 100ms must have elapsed before the read-out pulse is applied. The interval t_2 to t_3 should exceed 50ms.

Read-out by means of Positive-going and Negative-going Pulses

With this method the design value of the steady-state current need obey only Eq. 10, so that it may assume any value to the left of line (III) in the diagram of Fig. 21, that is less than 1·4mA in our example.

The design value of the current flowing during the positive-going pulse need obey only Eq. 10, which implies that this value must be situated to the right of line (I) in this graph, that is, it must exceed 1·25mA.

Finally, the design value of the current flowing during the negative-going pulse must comply with Eq. 5, so that this value must be situated to the left of line (II), and thus drop to 0·5mA or lower.

Fig. 22 again shows the unshaded area within the boundaries of which the design values of the current must lie. The steady-state value might be chosen to be 0·8mA (the preferred minimum value is 0·4mA) the positive-going pulse 2·5mA and the following negative-going pulse 0·25mA.

The duration t_1 to t_2 of the positive-going pulse should not be less than 100ms in this case. The interval t_2 to t_3 should exceed 50ms.

Manually operated Read-out Switch

The design value of the steady-state current need, in this case, obey only Eq. 2, which implies that its value must be situated to the right of line (I) in Fig. 23, and thus exceed 1·25mA in our example. When the read-out switch is operated, the design current must obey Eq. 5 for an instant; in other words it should drop below 0·5mA (without, however, dropping below the minimum permissible value of 0·05mA). Subsequently it is allowed to rise to a value which satisfies Eq. 10, that is to say, it may assume any value to the left of line (III).

The boundaries within which the design current must lie are shown in Fig. 24, where, for example, the steady-state current is chosen to be 2·0mA (recommended maximum value). When the switch is operated at t_2, 100ms after the multivibrators have reached their final state at t_1, the current drops to 0·2mA. After 50ms, at t_3, any wrongly conducting diode will be extinguished; the current then rises gradually up to 1mA. Once the circuit has been read out during the interval t_3 to t_4, the switch can be released so that the original condition is restored.

Effect of Read-out Circuit on Transistors

While the transistor multivibrator circuit and the diode read-out circuit are being designed some allowance may need to be made for the effects of the read-out circuit on the multivibrator. The following four effects are the only ones likely to be of any importance.

(a) If the 'wrong' read-out diode is temporarily conducting, the collector-emitter voltage of the corresponding (cut-off) transistor will be somewhat higher than would be the case in the absence of the read-out circuit.

(b) The base current of the saturated transistor will be slightly higher than it would in the absence of the read-out circuit.

(c) During the time that the 'correct' diode is conducting and its corresponding transistor is bottomed, the transistor collector current is increased by the diode current.

(d) In some sensitive forms of multivibrator circuit it is remotely possible that transient noise could pass from the diode circuit to the multivibrator and trigger the multivibrator, so leading to erroneous multivibrator states.

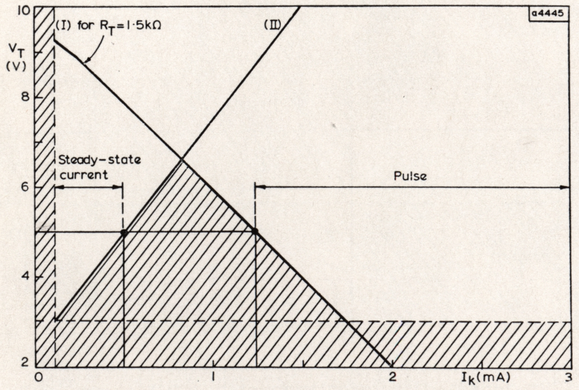


Fig. 17—Diagram in which curve (II) defines the maximum design value of the steady-state current and curve (I) the minimum design value of the current during the positive-going pulse if method (a) is chosen (at $V_T = 5\text{V}$ and $R_T = 1{\cdot}5\text{k}\Omega$)

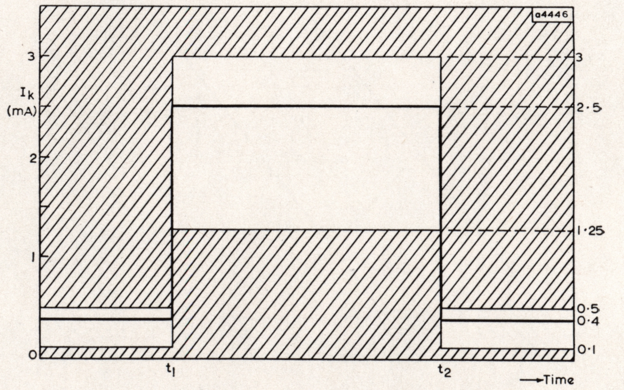


Fig. 18—Idealised representation of $I_k = f(t)$ corresponding to the design shown in Fig. 17. The circuit parameters should be so chosen that the design values of the current remain within the unshaded areas; it may assume the values indicated by the heavy line

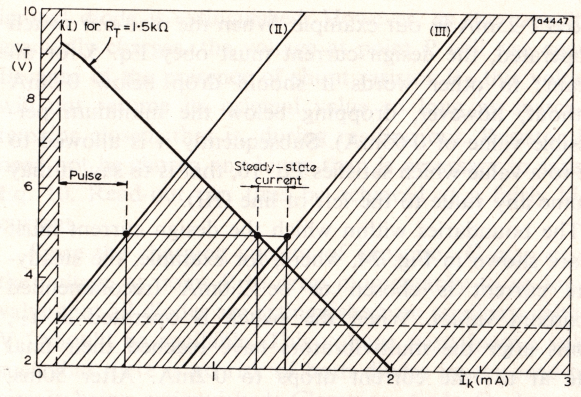


Fig. 19—Diagram in which curves (I) and (III) define the minimum and maximum design values of the steady-state current, and curve (II) the maximum design value of the current during the negative-going pulse if method (b) is chosen (at $V_T = 5V$ and $R_T = 1{\cdot}5k\Omega$)

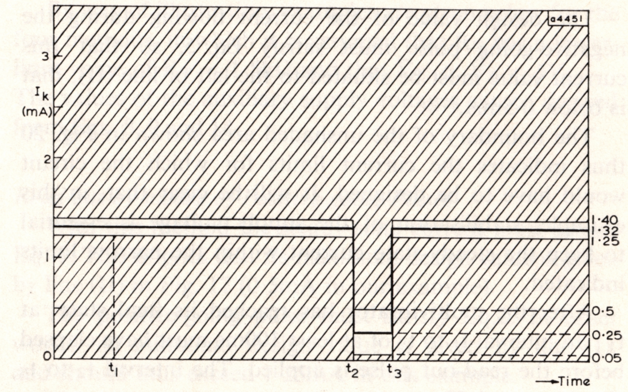


Fig. 20—Idealised representation of $I_k = f(t)$ corresponding to Fig. 19

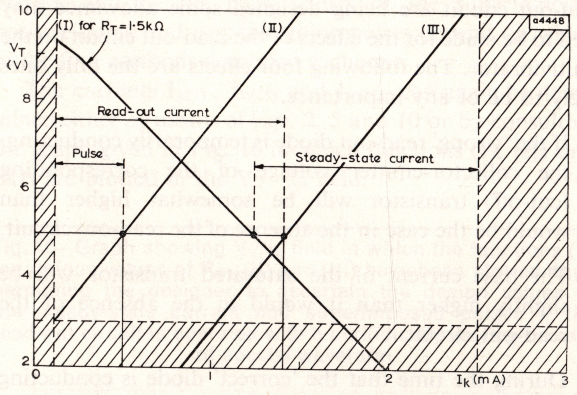


Fig. 21—Diagram in which curve (III) defines the maximum design value of the steady-state current, curve (II) the maximum design value of the current during the negative-going pulse, and curve (I) the minimum design value of the current during the positive-going pulse if method (c) is chosen (at $V_T = 5V$ and $R_T = 1{\cdot}5k\Omega$)

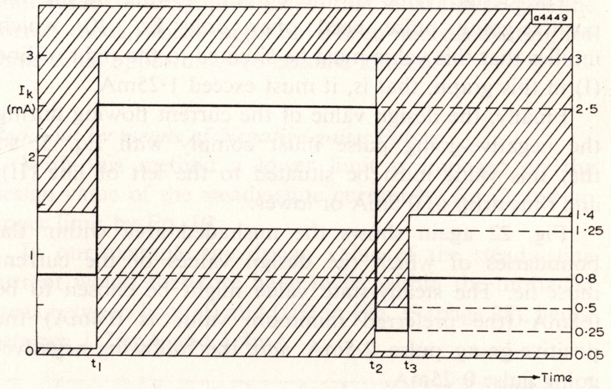


Fig. 22—Idealised representation of $I_k = f(t)$ corresponding to Fig. 21

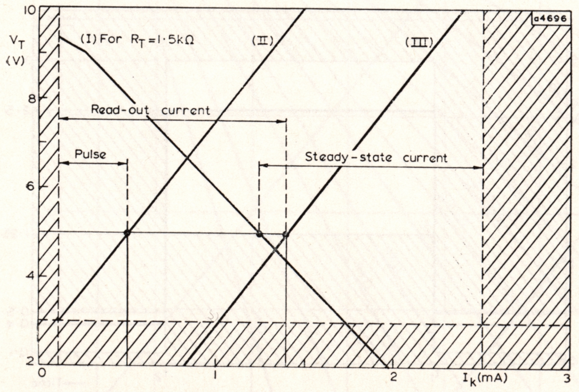


Fig. 23. Diagram in which curve (I) defines the minimum design value of the steady-state current, curve (II) the maximum design value of current during the negative-going pulse, and curve (III) the maximum design value of the read-out current if method (d) is chosen (at $V_T = 5V$ and $R_T = 1{\cdot}5k\Omega$)

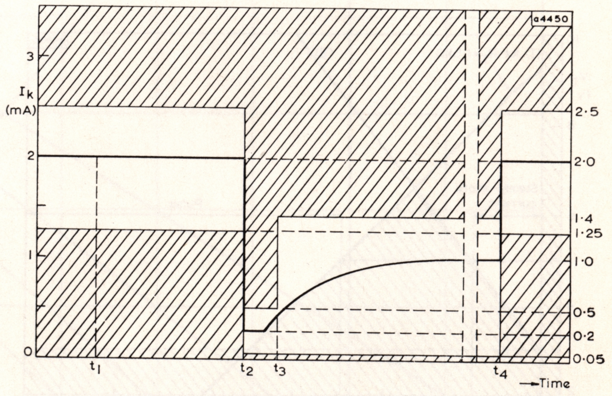


Fig. 24—Idealised representation of $I_k = f(t)$ corresponding to Fig. 23